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THE DENSITY OF STATES IN THE SEMICONDUCTORS WITH POSITION-DEPENDENT BAND STRUCTURE

BY

GH. ZET and MARIANA LATU

1. Introduction

If a semiconductor has a position-dependent lattice structure it also has a position-dependent band structure. In a semiconductor with a nonhomogeneous lattice structure the following are position-dependent: the bottom of the conduction band E_c, the top of the valence band E_v, the band gap E_g, the electron affinity χ and the density of states g. A graded composition of two or more components, a dependence on position of the impurity doping, a nonuniform strain or a nonuniform temperature also determine a position-dependent band structure.

J. Bardeen and W. Shokley [1] were the first to approach this problem. They have showed that E_c, E_v, E_g are position-dependent in a semiconductor with a varying elastic strain. P.R. Emtage [2] has considered the photovoltaic effect in semiconductors with position-dependent band gaps. S. Chattopadhyaya and V. Mathur [3] have considered the carrier distribution in graded band gap semiconductors under asymmetric band-edge gradients. T. Gora and F. Williams [4] have approached the theory of electronic states and transport in graded mixed semiconductors. A. Marshak and K.M. van Vliet have related the electrical current [5], the carrier densities [6], the law of the junction [7], the Shockley — like equations [8] and the effective density of states [9] in the semiconductors with position-dependent E_c, E_v, E_g, χ and g.

2. Generalized Density of States

In a nonhomogeneous semiconductor, the average electron density is [10]:

$$\bar{n} = \frac{1}{4\pi^3 V} \int_{(V)} \int_{(V_k)} f(\mathbf{k}, \mathbf{r}, T) \mathrm{d}\tau_k \, \mathrm{d}\tau_r , \tag{1}$$

where $f(\mathbf{k}, \mathbf{r}, T)$ is the electron distribution function. The distribution function in a equilibrium state of a semiconductor with a position-dependent band structure is [5]:

$$f^0(\mathbf{k}, \mathbf{r}, T) = f^0(E, E_{Fn}, T) = \{\exp [(E(\mathbf{k}, \mathbf{r}) - E_{Fn}(\mathbf{r}))/k_B T] + 1\}^{-1}, \tag{2}$$

where $E(\mathbf{k}, \mathbf{r})$ is the total energy of an electron and $E_{Fn}(\mathbf{r})$ — the quasi-Fermi level of the electrons. In relation (1), $\mathrm{d}\tau_k$ can be written in the form:

(3) $$d\tau_k = dS_E\, dk_\perp ,$$

where dS_E is an element of the constant energy surface in the space of the wave numbers and $dk_\perp$ is normal to the two constant energy surfaces, S_E and S_{E+dE}. In our case, the total energy of an electron depends on $\mathbf{k}$ and $\mathbf{r}$. Thus :

(4) $$dE = [\nabla_k E | dk_\perp + |\nabla_r E | dr_\perp ,$$

where $dr_\perp$ is normal to the constant energy surface in the real space. From relations (1), (3) and (4) we have :

(5) $$\bar{n} = \frac{1}{4\pi^3 V} \int_{(V)} \int_{(S_E)} \int_{(E)} f^0(E, E_{Fn}, T) \frac{dE - |\nabla_r E| dr_\perp}{|\nabla_k E|} dS_E d\tau_r .$$

The average electron density can be written also as a function of the electron density of states. In a semiconductor with the position-dependent band structure, the electron density of states depends on E and $\mathbf{r}$, $g(E, \mathbf{r})$. Thus $\bar{n}$ is given by :

(6) $$\bar{n} = \frac{1}{V} \int_{(E)} \int_{(V)} 2f^0(E, E_{Fn}, T) g(E, \mathbf{r}) dE\, d\tau_r .$$

From (5) and (6) we obtain :

(7) $$g(E, \mathbf{r}) = \frac{1}{8\pi^3} \int_{(S_E)} \frac{1 - |\nabla_r E| dr_\perp / dE}{|\nabla_k E|} dS_E .$$

In a homogeneous semiconductor we have $|\nabla_r E| = 0$ and then from relation (7) one obtain the well known relation :

(8) $$g_0(E) = \frac{1}{8\pi^3} \int_{(S_E)} \frac{dS_E}{|\nabla_k E|} .$$

If the total energy of an electron in a semiconductor with position-dependent band structure is the sum of the total energy of an electron in the nonperturbation crystal and of the perturbation part (this is for example, the case of those semiconductors having a nonuniform strain), i.e.

(9) $$E(\mathbf{k}, \mathbf{r}) = E_0(\mathbf{k}) + \delta E(\mathbf{r}),$$

then relation (7) may be written in the form :

(10) $$g(E, \mathbf{r}) = g_0(E) - \frac{1}{8\pi^3} \int_{(S_E)} \frac{|\nabla_r E|}{|\nabla_k E|} \cdot \frac{dr_\perp}{dE} dS_E .$$

3. Density of States in the Semiconductor having a Nonuniform Elastic Strain

First, we consider the case of the one-dimensional semiconductor having a nonuniform strain. We note the homogeneous lattice constant by a. We suppose that in nonhomogeneous crystals, the distance from the atom in the position x_{i-1} to the atom in the position x_i is $a+bx_i$.

Following the method of A. H. Marshak and K. M. van Vliet [5] we will consider a hypothetical homogeneous sample with the lattice constant $a+bx_i$. This homogeneous sample is obtained from the one-dimensional crystal having a nonuniform strain if the crystal structure between x_{i-1} and x_i is periodically repeated. The first Brillouin zone for this hypotetical homogeneous sample is $\left[-\frac{\pi}{a+bx_i}, \frac{\pi}{a+bx_i}\right]$.

If the one-dimensional semiconductor contains N atoms then, the wave number $\mathbf{k}$ can have, in the first Brillouin zone of the hypothetical homogeneous sample, the values:

$$k=0, \pm\frac{2\pi}{L_i}, \pm 2\frac{2\pi}{L_i}, \ldots, \pm N\frac{2\pi}{L_i}, \tag{11}$$

where $L_i=N(a+bx_i)$. Thus the size of the first Brillouin zone decreases with the increase of x. The values of $\mathbf{k}$ in the Brillouin zone decrease also.

The total energy of an electron in a one-dimensional semiconductor having a nonuniform elastic strain is [1]:

$$E(\mathbf{k}, x)=E_{c_0}+\frac{\hbar^2k^2}{2m^*}+C_1\Delta(x), \tag{12}$$

where E_{c_0} is the bottom of the conduction band in the homogeneous crystal and $\Delta(x)$ is the relative dilatation:

$$\Delta(x)=\frac{\delta l}{l}=\frac{bx}{a}. \tag{13}$$

From relations (7), (12) and (13) it follows that:

$$g(E, x)=\frac{k^2}{2\pi^2}\cdot\frac{1-C_1\frac{b}{a}\cdot\frac{1}{\frac{\hbar^2k}{m^*}\cdot\frac{\mathrm{d}k}{\mathrm{d}x}+C_1\frac{b}{a}}}{\frac{\hbar^2k}{m^*}}. \tag{14}$$

If N is very high, then the wave number k can be considered as a continuum function:

$$k=\xi\frac{2\pi}{a+bx}, \tag{15}$$

where $\xi\in[-1, +1]$. From relations (12), (14) and (15) we obtained the final relation:

$$(16)\quad g(E,x)=\frac{(m^*)^{3/2}(E-E_{c_0}-C_1bx/a)^{1/2}}{\sqrt{2}\pi^2\hbar^3}\left[1-\frac{1}{1-\dfrac{2^{3/2}\pi a\xi\hbar(E-E_{c_0}-C_1bx/a)^{1/2}}{C_1m^{*1/2}(a+bx)^2}}\right]$$

In a three-dimensional semiconductor, from relation (7) we can calculate the new density of state if we know the dispersion relation for the electrons and the relative dilatation.

In the case of a semiconductor having a nonuniform strain, relations (7) and (10) can be demonstrated also in the following way. If we denote by $\mathbf{r}_0$ the initial position of an atom and by $\mathbf{r}$, the position after deformation, the displacement vector is $\mathbf{R}=\mathbf{r}-\mathbf{r}_0$. We can expand $g(E,\mathbf{r})$ in a Taylor series :

$$(17)\quad g(E,\mathbf{R})=g(E,0)+\left.\frac{\partial g}{\partial\mathbf{R}}\right|_{\mathbf{R}=0}\mathbf{R}+\frac{1}{2}\cdot\left.\frac{\partial^2 g}{\partial\mathbf{R}^2}\right|_{\mathbf{R}=0}R^2+\ldots$$

Neglecting the high order terms we have :

$$(18)\quad g(E,\mathbf{r})=g_0(E)+C_g(\mathbf{r}-\mathbf{r}_0),$$

where :

$$(19)\quad C_g=\left.\frac{\partial g}{\partial\mathbf{R}}\right|_{\mathbf{R}=0}.$$

From (10) and (18) it follows that :

$$(20)\quad C_g(\mathbf{r}-\mathbf{r}_0)=-\frac{1}{8\pi^3}\int\limits_{(S_E)}\frac{|\nabla_{\mathbf{r}}E|}{|\nabla_{\mathbf{k}}E|}\cdot\frac{\mathrm{d}r_\perp}{\mathrm{d}E}\,\mathrm{d}S_E$$

and therefore we reobtain the formulae (7) and (10).

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DENSITATEA DE STĂRI ÎN SEMICONDUCTORI CU STRUCTURĂ A ZONELOR DE ENERGIE DEPENDENTĂ DE POZIŢIE

(Rezumat)

Se studiază densitatea de stări într-un semiconductor care prezintă o structură a zonelor energetice dependentă de poziţie. Se obţine a relaţie generală pentru densitatea de stări, dependentă de vectorul de undă **k** şi vectorul de poziţie **r**. Rezultatul obţinut se aplică la cazul unui semiconductor supus unei tensiuni elastice neuniforme.

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THE INFLUENCE OF THE OXIDE LAYER THICKNESS ON THE INTENSITY OF THE ELECTRIC CURRENT IN A SCHOTTKY CELL

BY

SOFIA MELINTE, ALEXANDRINA JEFLEA and N. MATEESCU

1. Introduction

1. The solar cells based on Schottky barrier represents a new alternative in performing the direct conversion of the solar energy into the electric one. These cells will be able to impose themselves in the future if their efficiency will exceed 10%, taking also into account their relatively simple fabrication technology, the relatively low amount of semiconducting materials used in the shape of thin films with thicknesses of the order of tens microns. The Schottky barriers of *n*-type, like Au/GaAs, Cr/Si, Ni/Si and CdS/Au, studied in [1]...[4], are well known. Cadmium sulfide is under the form of a thin film with polycrystalline structure.

Generally, for a Schottky barrier the charge distribution at the metal—semiconductor interface is given by the relation

$$Q_{ss}=Q_c \exp(-\Phi_B/E_0),$$

where Q_{ss} is the charge distribution at the interface; Q_c — a constant of integration; Φ_B — the height of the Shottky barrier and E_0 — a constant of distribution that can be temperature dependent in most of cases.

In order to discuss different parameters that characterize the Schottky barrier, it is necessary to remember the distribution of the energy and electric charge within this barrier. With these aim in view, in Fig. 1 the energetic diagram of a *n*-type Schottky barrier is represented for three cases: *a*—before the formation; *b* — during the formation; *c* — under different conditions of illumination, external electric field a.s.o.

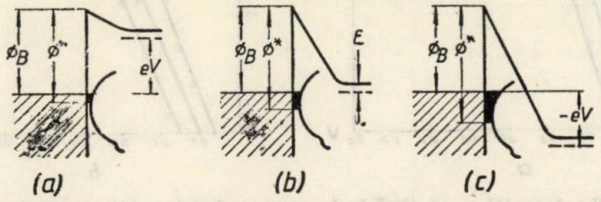

Fig. 1. — The relation between the barrier height Φ_B and the distribution of the electric charge at the metal — semiconductor interface.

The energy difference of the Fermi level will be

$$\Delta E_F = \Phi^* - \Phi_B.$$

The distribution of the surface energy states is given by the parameter Φ^* that increases with the application of a voltage or a illumination. One can also see that the Fermi level from the semiconductor tends to be equal to the Fermi level from the metal, while respecting the conduction zone border.

The current density J, of the Schottky barrier, is given by [7]

$$J = A^* T^2 \exp\left[-(\Phi_B - eV)/kT\right], \tag{1}$$

where A is a constant of integration, T — the absolute temperature, eV — the energetic difference between the Fermi level and the valence zone (Fig. 1), k — Boltzmann constant.

The relation (1) shows the dependence of the current density through a Schottky barrier on the temperature, barrier height and the position of the Fermi level in the semiconducting material.

2. Results and Discussion

The photovoltaic Schottky cells accomplished by the authors, being of the n-CdS/Cr and n-CdS/Au type, are in fact of the metal—insulator—semiconductor (MIS) type. On a copper substrate, CdS polycrystalline layers were obtained by vacuum flash evaporation ; above this layer a conducting Cu layer was deposited ; in order to protect the copper layer, a Cr layer was deposited to which the gold grid and contacts were attached. The second type of cell was made of CdS/Cr/oxide/Au, the grid and

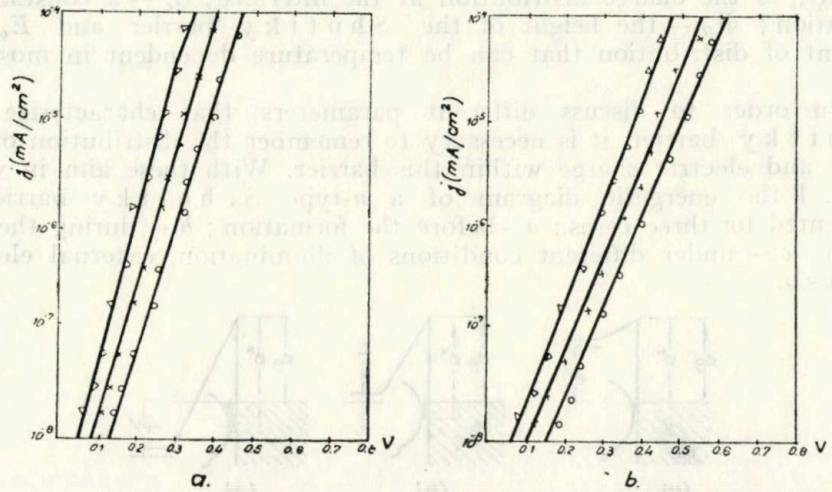

Fig. 2. — The logarithm of the dark current density j vs. the external voltage for different thicknesses of the oxide layer: a — CdS/Cr/oxide/Cu/Cr ; b — CdS/Cr/oxide/Au ; Δ — d=29,3Å : × — d=26,7Å ; o — d=30,2Å

the contacts. From the resistance and layer thickness measurements it was found that the Cr layer had a resistance $R = 900 \ldots 100\ \Omega/\square$ for a thickness of 50Å and that is why a conducting layer of Cu, Au or Pt is necessary. In order to obtain an efficient photovoltaic structure a severe control was necessary, concerning the execution of the oxide layer which, for thicknesses exceeding 50Å, becomes a perfect insulator, thus preventing the passage of the current through the corresponding structure.

In Fig. 2 one can see the influence of the oxide layer tikness on the current density for the two types of cells that we made : *a*—CdS/Cr/oxide Cu/Cr and *b*—CdS/Cr/oxide/Au.

From the two diagrams given in Fig. 2, one can notice that the value of the logarithm of the current density increases with the external voltage, but the slope of the two lines are different for the two types of cells. The current density maximum in the case *b* is reached for higher values of the applied voltage. This shows a decrease of the current density through the jonction when increasing the oxide layer thickness. Therefore the oxide layer thickness commands one of the basic parameters of the cell efficiency, namely the current intensity through the Schottky barrier.

3. Conclusion

In order to obtain highly efficient Schottky-type photovoltaic cells it is necessary, first of all, to obtain an oxide layer with a controlable thickness.

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INFLUENŢA GROSIMII STRATULUI DE OXID ASUPRA CURENTULUI ÎN CELULELE SOLARE BAZATE PE BARIERE SCHOTTKY

(Rezumat)

Se studiază două bariere Schottky de tip : CdS/Cr, CdS/Au, care au fost realizate prin evaporare în vid. După depunerea stratului de Cr, cu grosimea de ordinul angströmilor, prin tratamente termice în aer s-a obţinut stratul de oxid care tunelează curentul prin joncţiune. Se studiază efectul creşterii grosimii stratului de oxid asupra curentului prin barieră.

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D.C. 538.114

FERRIMAGNETIC MATERIALS USED IN MAGNETIC FLUX CONCENTRATORS

BY

C. PASNICU, SERGIU ISTRATE, EMIL LUCA and LAURENȚIU VARTOLOMEI

1. Introduction

In what follows, the preliminary results concerning the realisation of some ferritic cores used in the magnetic flux concentrators within the frequency range 300...500 KHz and at high magnetic fields are presented. The core of the concentrators is meant to improve the energy transfer from the primary to the secondary circuits.

The quality factor of a circuit is, principially, given by :

$$Q=\frac{W_t}{W_c}, \tag{1}$$

where W_c and W_t represent the consumed and, respectively, the transfered energy by the circuit.

In the case of a coil with a magnetic core, the quality factor is given by the relation :

$$Q=\frac{L'\omega}{R'}, \tag{2}$$

with $L'=L-L_0$; $R'=R-R_0$, where L, R, and L_0, R_0 represent the inductance and the resistance measured in the coil with core and air respectively, and $\omega=2\pi f$, f — the frequency of the signal applied to the coil.

R', representing the loss resistance, depends on the losses within the core, due to the hysterezis, eddy currents, magnetic viscosity, dimensional resonance and spin gyromagnetic resonance. As follows from the relation (2), a good quality factor implies an apparent permeability $\mu_a = L/L_0$ and a resistivity as high as possible, having in view that the loss resistance decreases with the decrease of the material electric conductivity.

2. The Experimental Method

The geometry of the core depends on the nature of its practical application ; for our purpose, it is necessary to use cylindrical bar samples.

The devices used for accomplishing the ferritic bars were : the ball mill, the die for bars, the press, and the furnace for thermal treatment.

The ball mill steel grinding enclosure and balls and is provided with a sealing, which allows the wet grinding.

The die is made of three separate parts, which permit to obtain crude bars with the following dimensions : $L=200$ mm and $\varnothing=12$ mm.

The compaction of the powder was made by means of a worm mechanical press provided with a limiting device. This limiting device was necessary because it was found that, beyond a certain value of the pressure, the crude samples stratify along a direction which is perpendicular to the pressure force. The samples were obtained by means of the method used in the most of ceramic techniques [1], [2].

The initial raw materials, namely the oxide in the case of the iron, and acetats, in the case of the other constituents of the mixed ferrite, have been wet grinded and homogenised, and mixture was presintered at 800°C. The bars obtained by pressing the presintered powder were subjected to a final treatment, performed for five hours at the temperature of 1 250°C, in the ambient atmosphere. The cooling was made slowly, together with the furnace.

The applications we have already mentioned impose the study of the behaviour of the ferritic cores under the conditions of high magnetizing fields ; that is why the measuring methods also will be specific [3], [4]. In order to simulate the intense magnetic fields, the primary circuit is supplied by a generator of current impulses, whose fundamental diagram is given in Fig. 1.

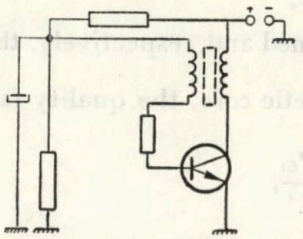

Fig. 1. — Generator of current impulses (pulse generator).

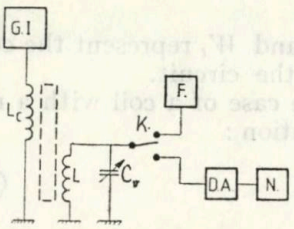

Fig. 2. — The schematic diagram used to determine the quality factor Q : GI — generator of current impulses ; F — frequency-meter ; DA — amplitude discriminator ; N — counter ; C_v — variable condenser ; K — switch.

The sample quality factor, Q, was determined by means of the diagram given in Fig. 2. The generator operating frequency is chosen to be at least ten times lower than the core operating frequency (500 KHz) and, by means of the variable condenser C_v, the secondary coil is turned on the measuring frequency (500 KHz). Then, by means of a universal counter the oscillations of the periodic damped signal are counted, for which the amplitude of the alternance is greater than V_0/e^{π}, where V_0 represents

the amplitude of the first impulse from the secondary circuit. The indicated figure will exactly be the value of the quality factor Q.

The apparent magnetic permeability μ_a is determined by introducing gradually the ferritic bar inside the generator coil (s. Fig. 3), until the frequency of 500 KHz is read in the secondary coil circuit. The introduced portion is measured and the value of the apparent permeability is determined, by comparison or by calculation.

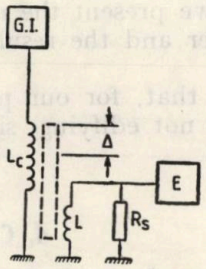

Fig. 3. — The diagram for determining the apparent magnetic permeability; GI — impulse generator; F — frequency-meter; R_s — load resistance; Δ — the distance on which the ferritic bar is introduced inside the generator coil.

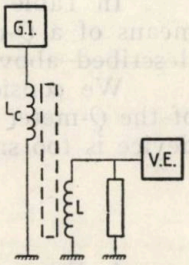

Fig. 4. — The diagram for the determination of the efficiency of magnetic flux concentration; VE — electronic voltmeter.

The efficiency of the energy transfer and, implicitely, the efficiency of concentrating the flux at high values of the magnetizing field, are determined by means of the diagram given in Fig. 4. By knowing the value of the induced voltage, read at the voltmeter, one can estimate the efficiency, η.

3. The Experimental Results

Two composition of mixed ferrites were realized, noted by A and B; the A ferrite contains Fe, Ni, Zn, Cu, and the B ferrite contains Fe, Ni, Zn, Cu, Ca.

The result of the measurements made on these samples were compared with those obtained on two ferrite bars, one of them being made by the Siemens Company, and noted by C, and the other, of Romanian production noted by D; the two bars are used in magnetic flux concentrators. The ferritic bars have the following dimensions; $(A) - L = 170$ mm; $\varnothing = 9$ mm; $(B) - L = 150$ mm; $\varnothing = 8$ mm; $(C) - L = 155$ mm; $\varnothing = 9$ mm; $(D) - L = 180$ mm; $\varnothing = 11$ mm.

The dimensions of the samples are necessary since the apparent magnetic properties of the materials depends on the geometry and the volume of magnetic core.

Table 1

The type of the ferrite	Q-meter			Specific methods				
	Q	f KHz	μ_a	Q	f KHz	Δ, [mm] $\mu_a = f(1/\Delta)$	U_{ind} V	η %
A	145	500	4.6	24	509	8	10.8	52
B	120	500	3.9	30	494	10	7.6	37
C	158	500	5.7	36	490	8	9.6	47
D	180	500	4.5	28	491	20	3.5	17

In Table 1, we present the results of the measurements obtained by means of a Q-meter and the results obtained by the measuring methods described above.

We consider that, for our purpose, the results obtained by means of the Q-meter are not edifying, since the magnetizing field given by this device is too small.

4. Conclusions

1. From the analysis of the results and taking into account the fact that the energy transfer depends on the factor of filling the coil with magnetic material, one can draw the conclusion that the C sample is the best, being closely followed by the B sample.

2. Both, the B and A samples are more convenient to be used, for the above mentioned purpose as compared to the D one.

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MATERIALE FERIMAGNETICE UTILIZATE ÎN CONCENTRATOARELE DE FLUX

(Rezumat)

Sînt prezentate şi comparate rezultatele obţinute pe patru tipuri de probe sinterizate, la care diferă compoziţia, două realizate de autori şi două utilizate în industrie, în vederea obţinerii miezurilor ferimagnetice necesare în concentratoarele de flux magnetic de mare putere.

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EFFET PELLICULAIRE DANS UN CONDUCTEUR CYLINDRIQUE FERROMAGNÉTIQUE, À FAIBLE NONLINÉARITÉ, EN RÉGIME QUASI-STATIONNAIRE HARMONIQUE

II. EFFET PELLICULAIRE FAIBLE

PAR

HUGO ROSMAN

1. Introduction

Dans une Note antérieure [1] on a étudié *l'effet pelliculaire net* dans un conducteur cylindrique ferromagnétique, à faible nonlinéarité, en regime quasistationnaire harmonique.

Le but de cette Note est d'étudier *l'effet pelliculaire faible* dans un conducteur cylindrique ferromagnétique, de section circulaire (Fig. 1), à faible nonlinearité, en régime quasistationnaire harmonique. Tout comme dans la Note précedente on accepte les hypothèses ci-dessous :

a) La caractéristique de magnétisation du matériel dont est confectionné le cylindre est approximée par le polynôme :

$$B=\mu_1 H-\mu_3 H^3,\ (\mu_1>0,\ \mu_3>0), \tag{1}$$

où B e H sont les valeurs efficaces de l'induction magnétique et de l'intensité du champ magnétique, qui oscillent en fonction du temps, selon ce qu'on suppose, d'après des lois harmoniques de pulsation ω.

b) La caractéristique de magnétisation est, en valeurs instantanées, linéaire, de sorte que le milieu puisse être considéré comme „inertiel" dans le sens précisé en [2].

c) Les hypothèses précédentes permettent à établir que la perméabilité magnetique du conducteur cylindrique est fonction de l'amplitude de l'intensité du champ magnétique, à savoir :

$$\mu=\mu_1-\frac{3}{4}\mu_3 H_m^2. \tag{2}$$

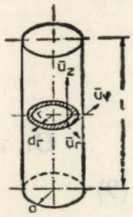

Fig. 1

Par opposition à la Note précédente on considère que l'effet pelliculaire, est faible, la profondeur de pénétration, δ, en vérifiant l'inégalité :

$$\delta \gg A, \tag{3}$$

où $A=\pi a^2$ représente l'aire de la section droite du conducteur cylindrique.

Il s'avère comme utile à préciser que le point de vue „inertiel", adopté dans l'étude de l'effet pelliculaire, offre un cadre avantageux pour l'effet

pelliculaire faible, en permettant à obtenir des résultats meilleurs dans le cas des fréquences basses. Sans doute, avec la restriction que les fréquences ne soient pas hausses le procédé proposé peut être utilisé même dans le cas de l'effet pelliculaire net, tout comme l'on a prouve dans [1], si la géometrie de la section droite du conducteur est convenable, de sorte que $\delta \ll A$.

On peut déterminer les composantes **E**, **H** du champ électromagnétique, dans le cas étudié dans ce qui suit, d'une façon assez simple et avec des approximations acceptables, en utilisant une méthode d'itération bien connue [3]. Cette méthode implique le nécessité de parcourir les étapes ci-dessous :

a) On admet, tout d'abord, que le courant est uniformément réparti dans la section du cylindre, de sorte que le densité du courant soit :

$$J_0 = \frac{I}{\pi a^2} \tag{4}$$

et par conséquent

$$\underline{\mathbf{E}}_0 = \frac{1}{\sigma}\underline{\mathbf{J}}_0, \quad \text{rot}\,\underline{\mathbf{H}}_0 = \sigma\underline{\mathbf{E}}_0, \tag{5}$$

où l'on a supposé que le conducteur est isotrope et homogène et le régime, quasistationnaire harmonique.

b) On détermine, successivement, des corrections $\mathbf{E}_k$, $\mathbf{H}_k$, ($k = 1, 2, \ldots, n$), par l'intégration des équations du champs électromagnétique :

$$\text{rot}\,\underline{\mathbf{E}}_k = -\mathrm{j}\omega\left(\mu_1 - \frac{3}{4}\mu_3 H^2_{m,k-1}\right)\underline{\mathbf{H}}_{k-1}, \quad \text{rot}\,\underline{\mathbf{H}}_k = \sigma\underline{\mathbf{E}}_k. \tag{6}$$

c) On obtient les valeurs finales des intensités du champs électrique et magnétique :

$$\underline{\mathbf{E}} = \sum_{k=0}^{n}\underline{\mathbf{E}}_k, \quad \underline{\mathbf{H}} = \sum_{k=0}^{n}\underline{\mathbf{H}}_k. \tag{7}$$

d) Les constantes d'intégration qui en résultent lors de l'intégration des équations (8) on les obtient en imposant les conditions :

$$\iint_{\Sigma}\underline{\mathbf{J}}_k\,\mathrm{d}\mathbf{A} = \sigma\iint_{\Sigma}\underline{\mathbf{E}}_k\,\mathrm{d}\mathbf{A} = 0, \quad H_{kt} = 0, \quad (k = 1, 2, \ldots, n). \tag{8}$$

2. Calcul de la première itération

On adopte un système de coordonnées cylindriques (Fig. 1), avec :

$$\underline{\mathbf{E}} = \mathbf{u}_z\underline{E}(r), \quad \underline{\mathbf{H}} = \mathbf{u}_\varphi\underline{H}(r), \quad \underline{\mathbf{J}} = \mathbf{u}_z\underline{J}(r). \tag{9}$$

Une fois déterminée la densité de courant J_0, à l'aide de la relation (4), la deuxième des équations (5) en étant, dans le système de coordonnées choisi, de la forme :

$$\frac{\partial}{\partial r}(r\underline{H}_0) = r\underline{J}_0, \tag{10}$$

le solution de celle-ci est [3] :

$$\underline{H}_0 = \frac{r}{2}\underline{J}_0. \tag{11}$$

On peut obtenir la première correction en intégrant l'équation (6_1)

$$\text{rot}\,\underline{\mathbf{E}}_1 = -j\omega\left(\mu_1 - \frac{3}{2}\mu_3 H_0^2\right)\mathbf{H}_0, \tag{12}$$

où l'on a tenu compte que $H_{m0}=\sqrt{2}\,H_0$. Avec une légère approximation il résulte, dans le système de coordonnées cylindrique, l'équation :

$$\frac{\partial}{\partial r}\underline{E}_1 = j\omega\left(\mu_1 - \frac{3}{8}\mu_3 r^2 \underline{J}_0^2\right)\frac{r}{2}\underline{J}_0 \tag{13}$$

et par conséquent

$$\underline{E}_1 = \frac{j\omega r^2}{4}\left(\mu_1 - \frac{3}{16}\mu_3 r^2 \underline{J}_0^2\right)\underline{J}_0 + \underline{C}_1'. \tag{14}$$

On détermine la constante d'intégration C_1' à l'aide de la première relation (8), à savoir :

$$\iint_{\Sigma}\underline{\mathbf{E}}_1\,d\mathbf{A} = \int_0^a\left[\frac{j\omega r^2}{4}\left(\mu_1 - \frac{3}{16}\mu_3 r^2\underline{J}_0^2\right)\underline{J}_0 + \underline{C}_1'\right]2\pi r\,dr = 0, \tag{15}$$

en obtenant

$$\underline{C}_1' = -\frac{j\omega}{8}a^2\left(\mu_1 - \frac{1}{8}\mu_3 a^2\underline{J}_0^2\right)\underline{J}_0 \tag{16}$$

et par suite la relation (14) devient :

$$\underline{E}_1 = \frac{j\omega\mu_1}{4}\left(r^2 - \frac{a^2}{2}\right)\underline{J}_0 - \frac{j\omega\mu_3}{64}(3r^4 - a^4)\underline{J}_0^3. \tag{17}$$

La deuxième des équations (6) permet à écrire :

$$\frac{1}{r}\cdot\frac{\partial}{\partial r}(r\underline{H}_1) = \frac{j\omega\mu_1\sigma}{4}\left(r^2 - \frac{a^2}{2}\right)\underline{J}_0 - \frac{j\omega\mu_3\sigma}{64}(3r^4 - a^4)\underline{J}_0^3 \tag{18}$$

qui, par intégration, conduit à :

$$\underline{H}_1 = \frac{j\omega\sigma r(r^2-a^2)}{16}\left[\mu_1 - \frac{1}{8}(r^2+a^2)\mu_3\underline{J}_0^2\right]\underline{J}_0 + \underline{C}_1''. \tag{19}$$

Compte tenu que $\underline{H}_1(a)=0$ il résulte

$$\underline{C}_1'' = 0 \tag{20}$$

et

$$\underline{H}_1 = \frac{j\omega\sigma r(r^2-a^2)}{16}\left[\mu_1 - \frac{1}{8}(r^2+a^2)\mu_3\underline{J}_0^2\right]\underline{J}_0. \tag{21}$$

3. Le champ électromagnétique à l'intérieur du cylindre ferromagnétique à faible nonlinéarité

Dans le cas de l'effet pelliculaire faible il est suffisant à prendre en considération seulement la première correction, de sorte que

$$\underline{E}_{nl}=\underline{E}_0+\underline{E}_1,\quad \underline{H}_{nl}=\underline{H}_0+\underline{H}_1. \tag{22}$$

Si l'on substitue, en (22), les expressions (5_1) et (17) de E_0 et E_1, respectivement (11) et (19) de H_0 et H_1, on obtient :

$$\underline{E}_{nl}=\frac{1}{\sigma}J_0+\frac{j\omega\mu_1}{4}\left(r^2-\frac{a^2}{2}\right)\underline{J}_0-\frac{j\omega\mu_3}{64}(3r^4-a^4)\underline{J}_0, \tag{23}$$

$$\underline{H}_{nl}=\frac{r}{2}J_0+\frac{j\omega\sigma}{16}r(r^2-a^2)\left[\mu_1-\frac{1}{8}(r^2+a^2)\mu_3\underline{J}_0^2\right]\underline{J}_0. \tag{24}$$

Si le milieu est linéaire et sa perméabilité magnétique est $\mu=\mu_1$, les intensités des champs électrique et magnétique sont données par :

$$\underline{E}_l=\frac{1}{\sigma}\underline{J}_0+\frac{j\omega\mu_1}{4}\left(r^2-\frac{a^2}{2}\right)\underline{J}_0, \tag{25}$$

$$\underline{H}_l=\frac{r}{2}\underline{J}_0+\frac{j\omega\mu_1\sigma}{16}r(r^2-a^2)\underline{J}_0. \tag{26}$$

On peut définir les écarts :

$$\Delta\underline{E}=\underline{E}_l-\underline{E}_{nl}=\frac{j\omega\mu_3}{64}(3r^4-a^4)\underline{J}_0^3, \tag{27}$$

$$\Delta\underline{H}=\underline{H}_l-\underline{H}_{nl}=\frac{j\omega\sigma\mu_3}{128}r(r^4-a^4)\underline{J}_0^3, \tag{28}$$

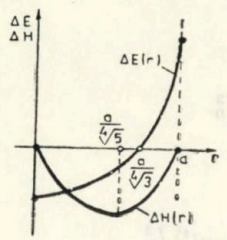

Fig. 2

dûs à la nonlinéarité de milieu. Les courbes $\Delta E(r)$, $\Delta H(r)$ sont représentées dans la Fig. 2. On observe que $\Delta E(r)$ est strictement croissante, en s'annulant pour $r=a/\sqrt[4]{3}$. Quant à l'écart $\Delta H(r)$ celui-ci s'annule aux extrémités de l'intervalle $[0, a]$ et admet un minimum pour $r=a/\sqrt[4]{5}$:

$$\Delta H_{\min}=-\frac{\omega\sigma\mu_3 a^5}{160\sqrt[4]{5}}. \tag{29}$$

Donc $\Delta E(r)\gtrless 0$ selon que $r\gtrless a/\sqrt[4]{5}$ tandis que $\Delta H(r)\leqslant 0$ pour n'importe quelle valeur $r\in[0, a]$. On constate que, dû à la nonlinéarité du milieu, l'intensité du champ magnétique, en conditions en reste égales, a une valeur plus réduite par rapport à celle qui correspond à un milieu linéaire.

4. Pertes de puissance

On peut déterminer la puissance active dissipée dans un conducteur ferromagnétique à faible nonlinéarité à l'aide de la relation

$$P_{nl}=\iiint_{v_{cond}}\frac{1}{\sigma}J_{nl}^2\,dv, \tag{30}$$

où

$$\underline{J}_{nl}=\sigma E_{nl}=\underline{J}_0+\frac{j\omega\mu_1\sigma}{4}\left(r^2-\frac{a^2}{2}\right)\underline{J}_0-\frac{j\omega\mu_3\sigma}{64}(3r^4-a^4)\underline{J}_0^3 . \tag{31}$$

Or

$$J_{nl}^2=J_0^2+\frac{\omega^2\sigma^2}{16}\left[\mu_1\left(r^2-\frac{a^2}{2}\right)-\frac{1}{16}\mu_3(3r^4-a^4)J_0^2\right]J_0^2=$$

$$=\left[1+\frac{\omega^2\sigma^2}{16}\mu_1^2\left(r^2-\frac{a^2}{2}\right)^2\right]J_0^2-\frac{\omega^2\sigma^2}{128}\mu_1\mu_3\left(r^2-\frac{a^2}{2}\right)(3r^4-a^4)J_0^4+ \tag{32}$$

$$+\frac{\omega^2\sigma^2}{4\,096}\mu_3^2(3r^4-a^4)^2J_0^6,$$

où l'on a considéré $\underline{J}_0=J_0e^{j0}$. Compte tenu que le conducteur est cylindrique :

$$dv=2\pi rl\,dr, \tag{33}$$

de sorte que l'expression (30) devient :

$$P_{nl}=\frac{2\pi l}{\sigma}\int_0^a\left\{\left[1+\frac{\omega^2\sigma^2}{16}\mu_1^2\left(r^2-\frac{a^2}{2}\right)^2\right]J_0^2-\right.$$

$$\left.-\frac{\omega^2\sigma^2}{128}\mu_1\mu_3\left(r^2-\frac{a^2}{2}\right)(3r^4-a^4)J_0^4+\frac{\omega^2\sigma^2}{4\,096}\mu_3^2(3r^4-a^4)^2J_0^6\right\}r\,dr. \tag{34}$$

Tous calculs effectués on obtient :

$$P_{nl}=\pi l\left[\left(\frac{a^2}{\sigma}+\frac{\omega^2\sigma}{192}\mu_1^2a^6\right)J_0^2-\frac{\omega^2\sigma}{512}\mu_1\mu_3a^8J_0^4+\frac{\omega^2\sigma}{5\,120}\mu_3^2a^{10}J_0^6\right]. \tag{35}$$

Si l'on tient compte de (4) on peut écrire la relation (35) sous la forme plus avantageuse :

$$P_{nl}=lI^2\left[\left(\frac{1}{\sigma}+\frac{\omega^2\sigma}{192}\mu_1^2a^4\right)\frac{1}{\pi a^2}-\frac{\omega^2\sigma}{512\pi^3}\mu_1\mu_3I^2+\frac{\omega^2\sigma}{5\,120\pi^5a^2}\mu_3^2I^4\right]. \tag{36}$$

Dans le cas des conducteurs cylindriques linéaires, en supposant $\mu=\mu_1$, il résulte :

$$P_l=lI^2\left(\frac{1}{\sigma}+\frac{\omega^2\sigma}{192}\mu_1^2a^4\right)\frac{1}{\pi a^2}. \tag{37}$$

Il est possible à définir dans ce cas un écart des pertes de puissance dû à la nonlinéarité du conducteur :

$$\Delta P=P_l-P_{nl}=\frac{\omega^2\sigma}{512\pi^3}\mu_3lI^4\left(\mu_1-\frac{\mu_3}{10\pi^2a^2}I^2\right). \tag{38}$$

On observe que l'écart ΔP peut être étudié comme fonction du courant I qui circule par le conducteur. La courbe $\Delta P(I)$ est représentée dans la fig. 3, où :

(39) $$I_m=\pi a\sqrt{\frac{20\mu_1}{3\mu_3}},\quad I_0=\pi a\sqrt{\frac{10\mu_1}{\mu_3}},$$

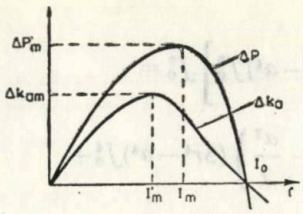

Fig. 3

avec $\Delta P>0$ pour $I<I_0$ et $\Delta P>0$ pour $I>I_0$. Évidemment le courant peut prendre des valeurs seulement dans l'intervalle dans le quel l'approximation (1) de la caractéristique de magnétisation est satisfaite, c'est-à-dire dans les champs pour les quels l'expression (1) est strictement croissante, à savoir pour

(40) $$H\in\left[0,\sqrt{\frac{\mu_1}{3\mu_3}}\right].$$

Si l'on introduit l'expression $H_{nl}=\sqrt{\mu_1/(3\mu_3)}$ dans la relation (24) on obtient une équation algébrique du troisieme degré en J_0 qui une fois solutionnée permet à déterminer, à l'aide de la relation (4), la valeur limite du courant, I, pour la quelle l'approximation de la caractéristique de magnétisation par la polynôme (1) est acceptable.

La valeur maximum de l'écart des pertes de puissance dû à la nonlinéarité du milieu résulte :

(41) $$\Delta P_m=\frac{25\pi}{864}\omega^2\sigma l a^4\frac{\mu_1^3}{\mu_3}.$$

5. Résistance en courant alternatif

Pour un conducteur cylindrique ferromagnétique à faible nonlinéarité on peut définir une résistance en courant alternatif :

(42) $$R_{canl}=l\left[\left(\frac{1}{\sigma}+\frac{\omega^2\sigma}{192}\mu_1^2a^4\right)\frac{1}{\pi a^2}-\frac{\omega^2\sigma}{512\pi^2}\mu_1\mu_3I^2+\frac{\omega^2\sigma}{5\,120\,\pi^5a^2}\mu_3^2I^4\right],$$

et, par conséquent, un facteur d'augmentation de la résistance en courant alternatif :

(43) $$k_{anl}=\frac{R_{canl}}{R_{cc}}.$$

Or $R_c=l/(\pi a^2\sigma)$ de sorte que

(44) $$k_{anl}=1+\frac{\omega^2\sigma^2}{192}\mu_1^2a^4-\frac{\omega^2\sigma^2}{512\pi^2}\mu_1\mu_3a^2I^2+\frac{\omega^2\sigma^2}{5\,120\pi^4}\mu_3^2I^4.$$

Dans le cas d'un conducteur linéaire, ayant $\mu=\mu_1$:

(45) $$k_{al}=1+\frac{\omega^2\sigma^2}{192}\mu_1^2a^4$$

et on peut définir un écart du facteur d'augmentation de la résistance en courant alternativ dû à la nonlinéarité du milieu :

$$\Delta k_a = k_{al} - k_{anl} = \frac{\omega^2\sigma^2}{512\pi^2}\,\mu_3 I^2\left(a^2\mu_1 - \frac{\mu_3}{10\pi^2}\,I^2\right). \tag{46}$$

On peut considérer l'écart Δk_a comme fonction du courant I ; le graphique $\Delta k_a(I)$ est représenté dans la fig. 3. Ici :

$$I'_m = \pi a \sqrt{\frac{5\mu_1}{\mu_3}} = \frac{\sqrt{3}}{2} I_m \tag{47}$$

et

$$\Delta k_{a\,m} = \frac{5}{1\,024}\,\omega^2\sigma^2 a^2 \mu_1^2. \tag{48}$$

Même dans ce cas, tout comme dans le paragraphe précedent, on doit tenir compte que le courant peut pendre des valeurs seulement dans l'intervalle dans le quel la caractéristique de magnetisation du matériel dont est confectionné le conducteur peut être approximée avec (1).

6. Conclusions

En utilisant une méthode d'itération on établit, dans les conditions d'un effet pelliculaire faible et pour un conducteur cylindrique ferromagnétique à faible nonlinéarité :

a) les expressions (23) et (24) des intensités du champ électrique et magnétique dans un point arbitraire situé à l'intérieur du conducteur ;

b) la relation (36) qui permet le calcul des pertes de puissance ;

c) la relation (44) qui permet à déterminer le facteur d'augmentation de la résistance en courant alternatif.

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Institut Polytechnique Jassy,
Chaire d'électrotechnique,
mesures et machines électriques

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EFECTUL PELICULAR ÎNTR-UN CONDUCTOR CILINDRIC FEROMAGNETIC CU NELINIARITATE SLABĂ, ÎN REGIM CVASISTAȚIONAR ARMONIC

II. Efectul pelicular slab

(Rezumat)

Folosind o metodă de iterație se stabilesc, în condițiile unui efect pelicular slab și pentru un conductor cilindric feromagnetic cu neliniaritate slabă :

a) expresiile (23) și (24) ale intensităților cîmpului electric și magnetic într-un punct arbitrar situat în interiorul conductorului ;

b) relația (36), de calcul a pierderilor de putere ;

c) relația (44), care permite determinarea factorului de creștere a rezistenței în curent alternativ.

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D.C. 62-52; 616.1

ON THE POSIBILITIES OF FAST FOURIER TRANSFORM APPLICATIONS IN HEART DISEASES DIAGNOSIS

BY

LIVIU GORAȘ, OCTAVIAN PETRUȘ, AKKRAM KHATIB, BORIS CONSTANTIN and IUSTINA ZAHARIA

1. Introduction

As it is well known, signal analysis consists of various techniques of expanding a given signal in a finite or infinite sum of orthogonal components [1] ... [3]. One reason for signal analysis is that it permits to compute easier the response of a linear system to a given input signal by summing the responses to each component. Another reason consists in the posibility of characterising the signal in the frequency domain, i.e. of being able to say something new about the initial signal looking at the weighting coefficients of the orthogonal components (spectrum).

This paper presents some aspect concerning this last reason applied to the study of signals taken from electrocardiograms.

2. Fourier Transform and Fast Fourier Transform (F.F.T.)

The Fourier transform represents a pair of signals $f(t)$ and $F(\omega)$ (spectral density of $f(t)$) related by :

$$(1) \qquad f(t)=\frac{1}{2\pi}\int_{-\infty}^{\infty}F(\omega)e^{-j\omega t}\,d\omega\,; \qquad F(\omega)=\int_{-\infty}^{\infty}f(t)e^{j\omega t}\,dt.$$

Obviously for real signals rigorous analytical calculations using (1) are generally impossible. The discrete approximation of (1) is represented by the discrete Fourier series (D.F.S.) which is now a pair of sequences of numbers related by :

$$(2) \qquad \begin{cases} \tilde{f}(kT)=\sum\limits_{k=0}^{N-1}\tilde{F}(n\omega_1)e^{-j\frac{2\pi}{N}kn}, & (k=0,1,\ldots,N-1), \\ \tilde{F}(n\omega_1)=-\frac{1}{N}\sum\limits_{k=0}^{N-1}\tilde{f}(kT)e^{j\frac{2\pi}{N}kn}, & (n=0,1,\ldots,N-1), \end{cases}$$

where

$$(3) \qquad \tilde{f}(kT)\approx f(kT), \quad \tilde{F}(n\omega_1)\approx F(n\omega_1), \quad (k=0,1,\ldots,N-1\,;\ n=0,1,\ldots,N-1)$$

and $\omega_1=2\pi/T$ is the sampling frequency, T — the signal period, $j=\sqrt{-1}$, N — the number of samples.

The errors implied by (3) depend on the shape of $f(t)$ and the number N of samples [2], [4].

We remark the algebraic character of (2), which makes it suitable for computer calculations. If we disregard the connexion between (1) and (2) we can write (2) in the form :

$$(4) \qquad a(k) = \sum_{k=0}^{N-1} A(n)\, e^{-j\frac{2\pi}{N}kn}, \quad A(n) = \frac{1}{N}\sum_{k=0}^{N-1} a(k) e^{j\frac{2\pi}{N}nk},$$

where $a(k)$ and $A(n)$ are sequences of numbers that form a discrete Fourier pair. The number $a(k)$ can eventually be obtained by sampling a given $f(t)$. It is clear that $A(n)$ are complex numbers conveying in their amplitude and phase the same information as $a(k)$ (and eventualy $f(t)$).

The term *Fast Fourier Transform* (F.F.T.) refers to certain algorithmes for computing the D.F.S. of a set of data $a(k)$. The advantage of using F.F.T. consists in the computational time saving compared with the classical algorithm [3]...[5].

From (4) becomes also evident the fact that the sequences $a(k)$ and $A(n)$ are periodic with period N, it follows that only the samples of one period are of interest.

3. Computer Program Description

For the calculation of amplitude and phase spectrum of a signal given in a discrete form we realised a Fortran computer program in which we have used as „building blocks" some high quality routines from NCAR Computer Soubroutine Library [6]. More specifically we used the routines FOUR2, BITRV, COOL2 and FIXRL. These routines represent the NCAR portable version of Norman Brenner implementation (MIT Lincoln Laboratory) of original Cooley — Tucky algorithm [6].

We created and validated a double precision version of these routines for roumain computer Felix C-256 from „Al. I. Cuza" University, Computing Center. The program reads signal data represented as real variables, transforms them in complex variables, performs Fourier transform, calculated and writes in a suitable form the amplitude and phase spectrum.

The program was carrefully tested with some problems with easy numbers worked using paper and pencil.

4. F.F.T. of Electrocardiogram Signals. Numerical Results

An electrodiagram consists of a number of rather periodic signals recorded from different parts of the human body, their shape allowing the physician to draw certain conclusions concerning the heart behaviour. Taking into consideration the almost periodic character of the heart signals, we sampled one period of each signal with $N=64$ samples (this number seems to be sufficient for our purposes). We computed the amplitudes and the phases of the corresponding $A(n)$ as to our knowledge such a treatement hasn't been reported so far.

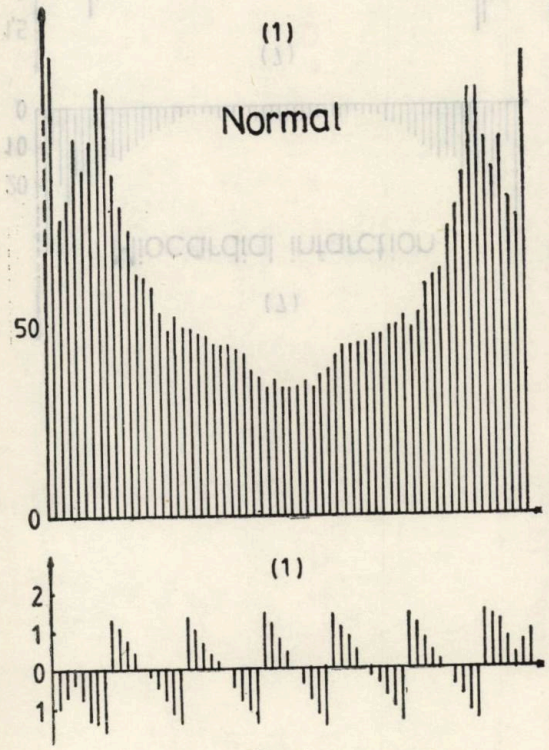

Fig. 1

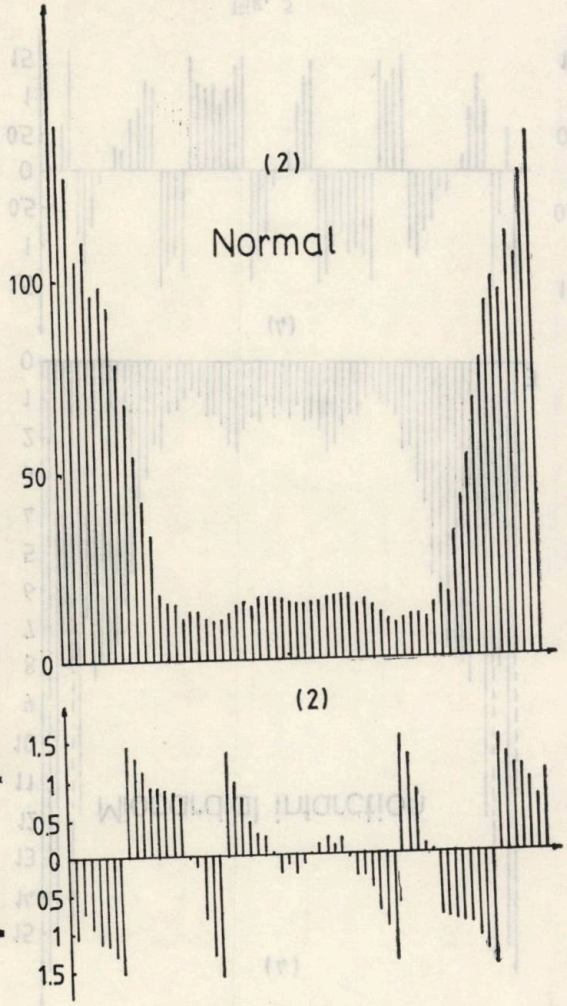

Fig. 2

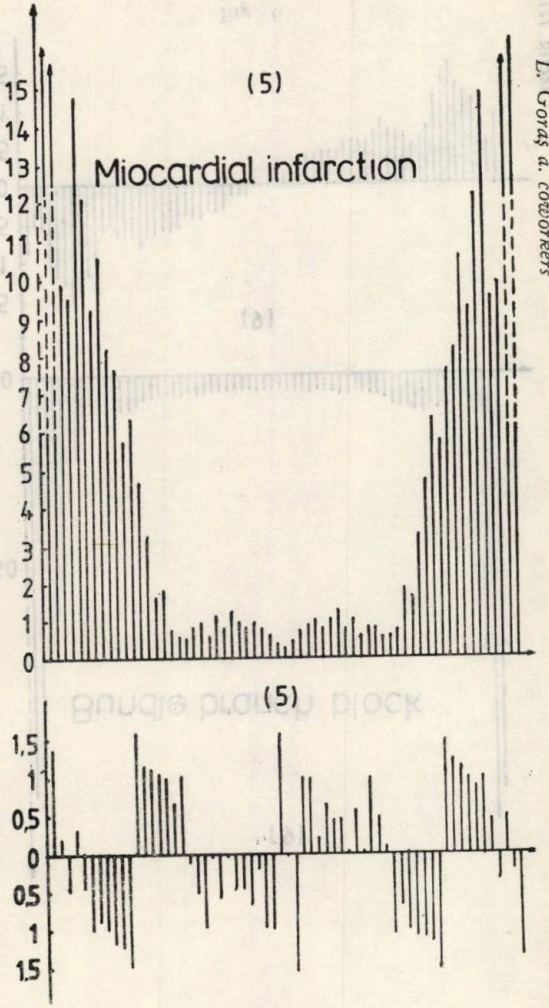

Fig. 3

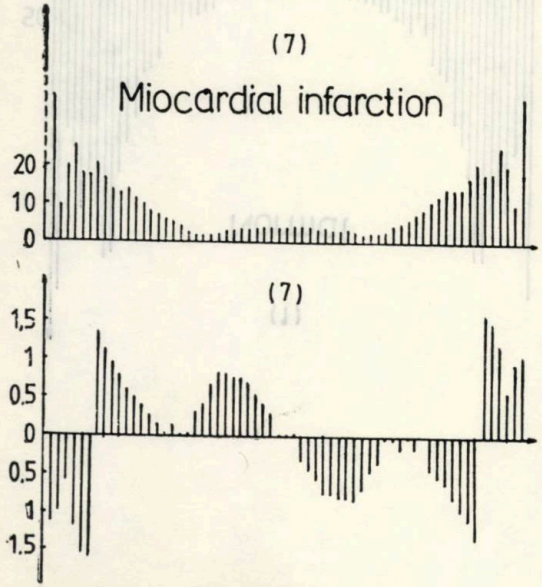

Fig. 4

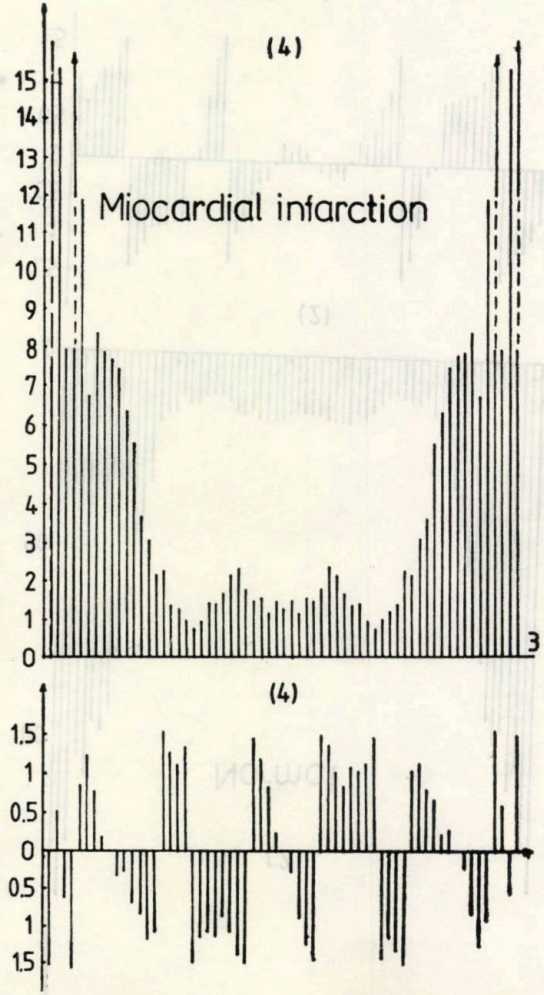

Fig. 5

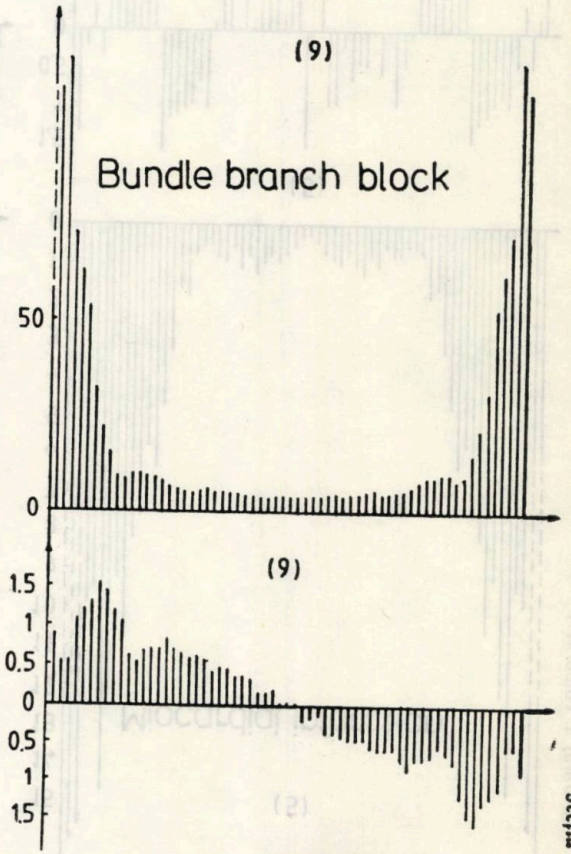

Fig. 6

The electrocardiogram signals, selected by a physician (Prof. Dr. N e g o i ț ă) from the Hospital nr. 1, Jassy, and their spectrum are presented in Fig. 1...6. In the first two figures the time domain and the amplitude and phase components of the spectrum for two healthy patients are shown. The next four figures show signals from ill pacients as follows : Fig. 3...5 correspond to patients having miocardial infarction and Fig. 6, to a pacient having bundle branch block.

At a first glance we can observe certain differences in the phase variations between healthy and ill patients. Further investigations, eventually with a greater N, could reveal some other peculiar features. Of course the time domain of the signals remains as important as before but we can expect an improvement of the investigation by means of the frequency domain.

5. Concluding Remarks

It seems that the F.F.T. techniques could be helpful for heart desease diagnosis if the physician could make connections between the shape of the spectrum and the heart desease.

Most probably, the conclusions commonly drawn by means of the time domain signals could be enlarged if aspects from the frequency domain are taken into consideration too. We suggest that these aspects could be : amplitude and phase shape, ratio between amplitudes or phase of different harmonics etc.

By means of this technique it could be also possible to find new criteria in heart deseases diagnosis and to reveal some new aspects of the heart activity.

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POSIBILITĂȚI DE APLICARE A TRANSFORMATEI F O U R I E R RAPIDE (T.F.R.) ÎN DIAGNOSTICAREA BOLILOR DE INIMĂ

(Rezumat)

Se prezintă cîteva aspecte ale utilizării transformatei F o u r i e r rapide aplicate la studiul semnalelor de pe electrocardiograme. Se efectuează analiza unor semnale de pe electrocardiograme pentru pacienți sănătoși și bolnavi, obținîndu-se spectrul în amplitudine și fază. Se remarcă faptul că tehnica T.F.R. poate fi folosită în diagnosticare unor boli de inimă, accentul punîndu-se pe spectrul de frecvență al fazei.

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C.Z. 621.317.725.082.72

UN VOLTMETRU ELECTROSTATIC CU TREI GAME COMUTABILE SUB TENSIUNE

DE

M. ANTONIU și EMIL VREMERĂ

1. Introducere

Într-o lucrare anterioară [1] au fost prezentate două variante de voltmetre electrostatice (cu ac și cu indicator optic) de 7,5/15/30 kV, la care gamele pot fi schimbate și sub tensiune; la ambele s-au folosit electrozi activi cu geometrie cilindrică-sferică. Cercetările noastre ulterioare au arătat că, pentru tensiuni nominale pînă la 10 ... 15 kV, se pot folosi și electrozi plani-paraleli, a căror tehnologie e mai simplă și mai ieftină.

În cele ce urmează se prezintă două variante de voltmetre cu electrozi plani-paraleli, cu trei game de tensiune. Asemenea aparate sînt foarte utile la măsurări de laborator în domeniul tehnicii tensiunilor înalte.

2. Varianta cu electrod mobil simplu

2.1. Descriere

Schema de principiu este dată în fig. 1 și 2. Modificarea gamelor (tensiuni nominale): U_n, U'_n, U''_n se realizează prin modificarea distanței (d) dintre electrozii activi (fig. 1), cu ajutorul unui sistem de trei electrozi de tensiune (*1*, *1'*, *1''*) fixați pe un ax electroizolant (*2*) cu ajutorul unei bucșe de bronz (*3*); prin rotirea acestui ax, cu cheia (*4*), electrozii de tensiune sînt aduși, unul după altul, în zona de lucru a electrodului mobil (*5*), conectat la borna de masă. Tensiunea de măsurat, U, aplicată între aceasta din urmă și borna izolată (*7*), este transmisă la electrozii de tensiune printr-un palpator metalic (*8*) care realizează contactul cu bucșa (*3*). Rezistența de protecție R_p servește la limitarea curentului în resortul antagonist (*9*) în caz de scurtcircuit (străpungere) între electrozii activi, iar magnetul M, la amortizarea organului mobil.

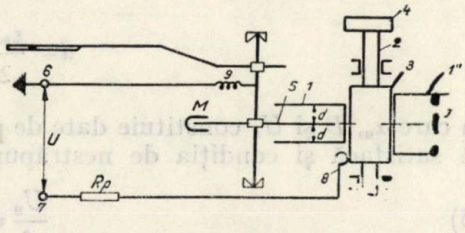

Fig. 1

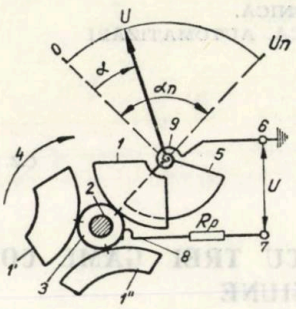

Fig. 2

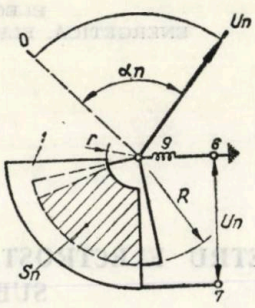

Fig. 3

2.2. Elemente de calcul

Neglijînd influența electrozilor *l'*, *l''* și a bucșei (*3*), mecanismul de măsură din fig. 2 devine un mecanism obișnuit pătratic (fig. 3) cu ecuația de funcționare

$$\alpha = \frac{KU^2}{2D}, \tag{1}$$

care pentru $\alpha = \alpha_n$ (capăt de scară) devine :

$$\alpha_n = \frac{KU_n^2}{2D}, \tag{2}$$

unde :

$$K = \frac{\varepsilon_0 S_n}{d}$$

și în care S_n reprezintă suprafața electrodului mobil care avansează între plăcile (cadranele) electrodului de tensiune (*1*) la variația deviației de la $\alpha = 0$ la $\alpha = \alpha_n$ (fig. 3). ε_0 — permitivitatea aerului, d — distanța de lucru (întrefierul) dintre electrozii activi, iar D — cuplul specific al resortului antagonist (*9*). De aici rezultă o primă relație de calcul :

$$d = \frac{\varepsilon_0 S_n U_n^2}{2\alpha_n D},$$

în care α_n, D și U_n constituie date de proiectare. Distanța de lucru d trebuie să satisfacă și condiția de nestrăpungere :

$$\frac{U_n}{d} \leqslant E, \tag{5}$$

în care E reprezintă intensitatea cîmpului electric dintre electrozii activi și care, pentru mecanisme pe ax cu lagăre (cazul de față), se alege $E = 0{,}3 \ldots 0{,}6$ kV/mm [2]. Condiția (5) aplicată și electrozilor *l'* și *l''* conduce la o a doua relație de calcul :

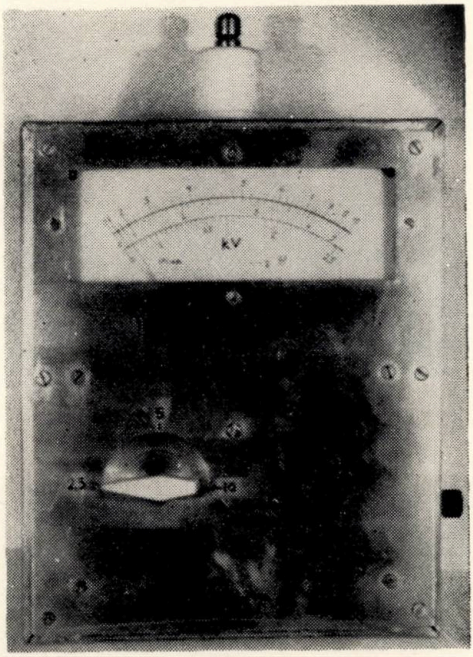

Fig. 4

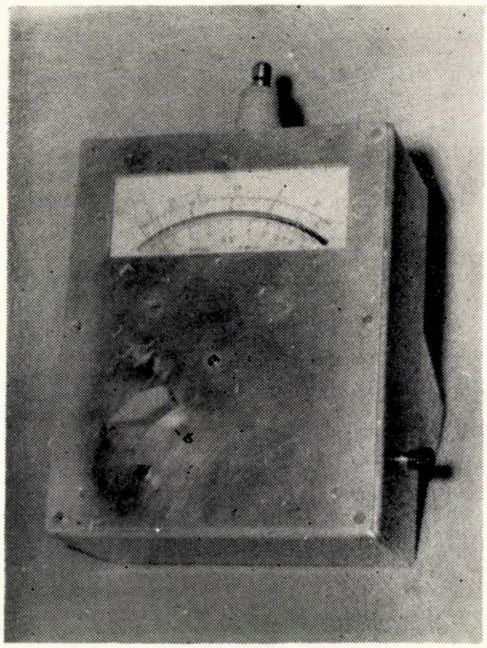

Fig. 6

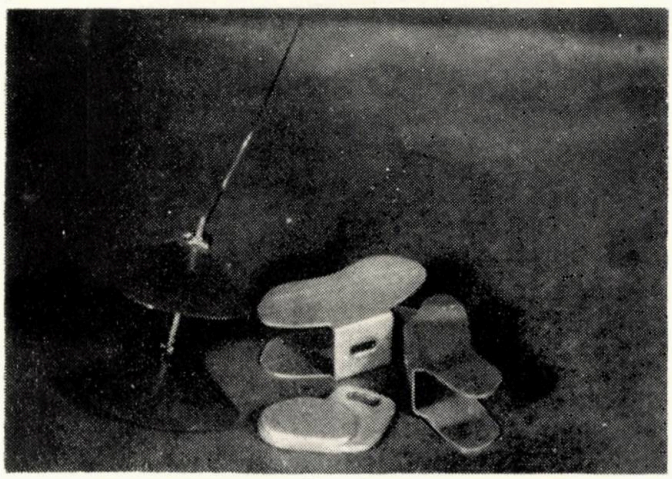

Fig. 7

$$\frac{U_n}{d} = \frac{U'_n}{d'} = \frac{U''_n}{d''}. \tag{6}$$

În fine, substituind pe (5) în (4), pentru egalitate se obține relația

$$U_n = \frac{2\alpha_n D}{\varepsilon_0 E S_n}, \tag{7}$$

care arată că la schimbarea gamelor (U_n), în afară de modificarea lui d trebuie de modificat cel puțin unul din parametrii E, S_n deoarece α_n și D trebuie să rămînă constanți. Încercările experimentale ne-au condus la concluzia că e mai avantajos de modificat S_n.

Observație. Calculele de mai sus au fost efectuate neglijînd influența celorlalți doi electrozi de tensiune (l' și l'') În realitate această influență nu e total neglijabilă și se răsfrînge în principal asupra formei scării: electrodul din dreapta (l'' — fig. 2) produce un cuplu parazit de același semn cu cel activ, iar cel din stînga (l'), acționînd asupra acului indicator, produce un cuplu de semn opus care provoacă îndesirea diviziunilor în partea finală a scării. Acest ultim aspect este mai evident pe gamele U'_n și U''_n și poate fi ameliorat cu ajutorul unui ecran.

2.3. Exemplu de calcul și construcție

Fie de realizat un voltmetru cu organ mobil pe lagăre și trei game: $U_n = 2{,}5$ kV, $U'_n = 5$ kV și $U''_n = 10$ kV.

Din considerente de proiectare [2] se adoptă: $\alpha = 80° \approx 1{,}4$ rad, $D = 4$ μNm/rad și $E = 0{,}5$ kV/mm. Cu ajutorul acestor date, din (6) se obțin: $d = 2{,}5/0{,}5 = 5$ mm; $d' = 10$ mm și $d'' = 20$ mm, iar din (4), transcrisă și pentru electrozii l' și l'', rezultă ($\varepsilon_0 = 8{,}89 \cdot 10^{-12}$ F/m): $S_n \approx 10$ cm², $S'_n = 5$ cm², $S''_n = 2{,}5$ cm². În fine, din fig. 3 rezultă relația:

$$S_n = \frac{\pi(R^2 - r^2)\alpha_n}{2} \tag{8}$$

din care se obține ($r^2 \ll R^2$):

$$R = \sqrt{\frac{2S_n}{\alpha_n}} = \sqrt{\frac{2 \cdot 10}{1{,}4}} \approx 35 \text{ mm}.$$

Cum R este același pentru toți electrozii de tensiune, mai rămîn de calculat r' și r'', din relația (8), transcrisă pentru electrozii l' și l''.

Voltmetrul construit pe baza acestor date este arătat în fig. 4. Se remarcă fenomenul îndesirii diviziunilor la fine pe scările de 5 și de 10 kV.

3. Varianta cu electrod mobil dublu

Varianta cu electrod mobil simplu conduce la un organ mobil ușor, însă, la tensiuni nominale de peste 7 ... 10 kV, crește mult dimensiunea axială a aparatului, care trebuie să fie cel puțin $4d''$. La varianta cu electrod mobil dublu, electrozii de tensiune fiind dispuși între plăcile electrodului mobil (conectat la masă), această dimensiune se reduce aproape la jumătate ($2d''$), însă organul mobil rezultă cu 30 ... 50% mai greu și deci trebuie

prevăzut un cuplu specific D, corespunzător, mai mare, pentru a nu creşte eroarea de frecare în lagăre.

Schema de principiu a variantei cu electrod mobil dublu este arătată în fig. 5. Funcţionarea este aceeaşi ca şi la cea din fig. 1, cu deosebirea că, aici, electrodul mobil îmbracă electrodul de tensiune. Metodica de calcul rămîne aceeaşi ca mai înainte.

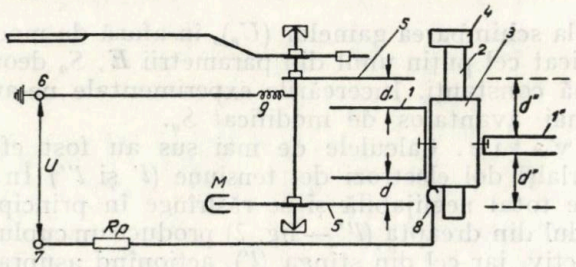

Fig. 5

În fig. 6 se reprezintă un voltmetru de 4/8/15 kV iar în fig. 7, electrozii de tensiune şi organul mobil. Aparatul se încadrează în clasa 1,5. Alţi parametri : lungimea scării gradate 120 ... 150 mm, rezistenţa de intrare $>10^{12}\ \Omega$ capacitatea de intrare <18 pF, dimensiuni : $290 \times 230 \times 140$ mm, masă ≈ 3 kg.

4. Concluzii

Utilizînd electrozi plani-paraleli se pot realiza voltmetre electrostatice cu trei game ce pot fi comutate şi sub tensiune, ceea ce constituie un mare avantaj pentru măsurările în domeniul tensiunilor înalte. Asemenea aparate se pot realiza pentru tensiuni nominale de 1 ... 15 kV şi precizii 1 ... 2,5 din lungimea scării gradate.

Primită la 24 martie 1982

Institutul politehnic, Iaşi,
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VOLTMÈTRE ÉLECTROSTATIQUE À TROIS GAMMES COMMUTABLES SOUS TENSION

(Résumé)

On présente deux voltmètres électrostatiques à aiguille indicatrice, de 2,5/5/10 KV et, respectivement 4/8/15 KV, gammes qui peuvent être changées même lorsque l'appareil est sous tension. Le changement des gammes est effectué par la modification de la distance entre les électrodes actifs.

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D.K. 536,5

ÜBER DIE MESSUNG DER DURCH WARMES WASSER TRANSPORTIERTEN WÄRMEMENGE

VON

M. CREŢU und C. HARJA

1. Einführung

So wie andere Energieformen, wird auch die Wärme dem Nutzer mittels eines Wärmetragers geliefert der meist für technologische Erwärmungen oder Raumheizung aus Dampf oder warmem Wasser besteht.

Als Änderung der inneren Energie des Wärmetragers ist die zurückbehaltene Wärmemenge von seiten des Verbrauchers proportional der Masse m und der Temperaturänderung des Wärmetragers, $\Delta\theta$, wobei der Proportionalitätsfaktor der spezifischen Warmekapazität, c, entspricht:

$$\Delta q = cm\Delta\theta. \tag{1}$$

Die obengenannte Abhängigkeit ist ausschließlich nur für kleine Temperaturdifferenzen gültig, denn die spezifische Wärme des Wärmetragers ist von der Temperatur abhängig. Gleichfalls ist zu erwähnen, daß in der Mehrzahl der Fälle die spezifische Wärme bei konstantem Druck, c_p, eine Rolle spielt, wobei die Masse durch das Volumen, eine leicht zu messende Größe, und der Dichte, ein ebenfalls von der Temperatur abhängiger Parameter, bestimmt wird [4], [5].

Wenn im Temperaturintervall zwischen θ_1, für die Flüssigkeit in der Hinleitung und θ_2, für die Flüssigkeit in der Rücklaufleitung der Heizungsanlage nicht ein Phasenwechsel stattfindet, dann kann die transportierte Wärmemenge durch die Beziehung :

$$\Delta q = V \int_{\theta_2}^{\theta_1} \rho(\theta) c_p(\theta)\, d\theta$$

wiedergegeben werden.

Weil in der Praxis das Integral des Produkts $\rho(\theta)c_p(\theta)$ über einen gegebenen Temperaturbereich schwierig zu verwirklichen ist, sind folgende Vereinfachungen von Nutzen :

a) Man benutzt den Mittelwert für die spezifische Wärme bei konstantem Druck :

$$c_{p_m} = \frac{1}{\theta_1 - \theta_2} \int_{\theta_2}^{\theta_1} c_p(\theta)\, d\theta. \tag{3}$$

b) Man mißt das Volumen des Wärmetragers in der Hinleitung, wobei im Druckbereich 1 ... 20 Bar die Einwirkungen der Druckänderungen auf die Flüssigkeitsparameter vernachläßigt werden können.

Mit diesen Vereinfachungen erhält man die Gleichung

$$q = V_{\theta_1} \rho_{\theta_1} c_{p_m} (\theta_1 - \theta_2) \tag{4}$$

und somit evident eine Fehlergrenze, die nicht unterschritten werden kann.

Die laufende Bestimmung der Wärmemenge wird technisch durch die Messung der Durchflußwärmemenge q durchgeführt, das heißt durch Messung der gelieferten Wärmemenge in einer Zeiteinheit und Integration oder Aufsummierung aller der Meßdauer entsprechenden Momentwerte q, bzw. der Beziehungen:

$$\dot{q} = \dot{V}_{\theta_1 \rho_{\theta_1}} c_{p_m} (\theta_1 - \theta_2), \qquad (5)$$

$$q = \int_{t_1}^{t_2} \dot{V}_{\theta_1 \rho_{\theta_1}} c_{p_m} (\theta_1 - \theta_2) dt, \qquad (6)$$

in denen t_1 und t_2 die Anfangs- und Endzeiten des Meßprozeß darstellen [4], [10].

Wenn einer der Parameter, die Durchflußmenge $\dot{V}$ oder die Temperaturdifferenz, $\theta_1 - \theta_2$ als konstant während der Messung angenommen werden kann, dann kann die Wärmemenge nur ungefähr mit Hilfe einer Wasseruhr oder dem Differenzthermometer bestimmt werden [3], [8].

2. Betrachtungen zum Aufbau der Wärmezählers bei Warmwasser Wärmeträger

Die Wärmezähler, bei dem der Wärmeträger außchließlich aus warmem Wasser besteht, arbeitet auf dem Prinzip der Integration über die Zeit des Produktes aus Durchflußvolumen der Flüssigkeit und Temperaturdifferenz.

Da jedoch die Wärmemenge durch den mengenmässigen Durchfluß bestimmt wird, enthält der Zahler, um genaue Messungen dürchführen zu können, Blöcke zur Ausgleichung der Schwankungen der von der Temperatur abhängigen Dichte und spezifischen Wärme und manchmal auch Anzeigeinstrumente für den Durchfluß und die Temperaturdifferenz [1], [2], [5], [6].

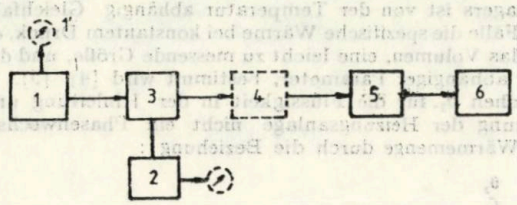

Abb. 1

In Abb. 1 wird schematisch ein Wärmezähler vorgestellt mit folgenden Bedeutungen: *1* — Durchflußmengenumwandler, der wahlweise an einen Durchflußmengenzähler angeschlossen sein kann; *2* — Umwandlerblock der Differenzen zwischen Hin- und Rücklauftemperatur, der wahlweise mit Anzeigegerät *2'* vorgesehen ist; *3* — Block, durch den man das Produckt erhält; *4* — Kompensationsblock (wahlweise), *5* — Integrierer, *6* — Block für die Anzeige des Resultats.

Die Literatur weist zur Zeit auf zwei Konstruktionstypen von Wärmemengenzählern hin: der mechanische und der elektrische Typ [3], [4], [8], [9]. Aber den Fortschritt bei der Konstruktion von Durchflußmessern berücksichtigend und besonder die Entwicklung von elektromagnetischen Durchflußmessern in letzter Zeit, betrachten wir die folgende Klassifikation als angebracht, die, nach unserer Meinung, das gegenwärtige Stadium und die Perspektive der Technik in diesem Bereich besser widerspiegelt:

a) mechanische Wärmezähler, bei denen alle Baugruppen mechanischer Natur sind;

b) elektromechanische Wärmemengenzähler, die in ihrem Aufbau sowohl mechanische als auch elektrische Baugruppen beinhalten;

c) elektrische Wärmemengenzähler, die keine beweglichen mechanischen Elemente enthalten, wobei die primären Signale der Durchflußmenge oder der Temperatur analog elektronisch oder numerisch bearbeitet werden.

Bei den mechanischen Wärmemengenzählern, entsprechend der oben erwahnten Klassifikation, wird das Produkt aus Durchfluß und Temperaturdifferenz durch Mechanismen, bestehend aus Nocken und Profilgetrieben, gekoppelt mit einem mechanischen Integrierer, gebildet, während bei den elektromechanischen Wärmemengenmessern das Produkt auf mechanischem oder analog elektrischem Wege erhalten werden kann, wobei der Integrierer gewöhnlich elektromechanischer Natur ist.

Die mit gesteuertem Kontakt versehenen W o l t m a n Durchflußmesser können als Grundlage für die Ausführung von elektromechanischen Zählern mit digitaler Primärsignalverarbeitung dienen.

3. Elektrische Wärmemengenzähler

Sie basieren ausschließlich auf Durchflußmengenumwandlern ohne bewegliche Elemente, wobei als Temperaturumwandler Thermowiderstände oder Thermoelemente verwendet werden. In diesem Sinne hebt sich der Wärmemengenzähler mit einem elektromagnetischen Durchflußmengenumwandler ab, der auf Basis einer der drei folgenden Varianten funktionieren kann.

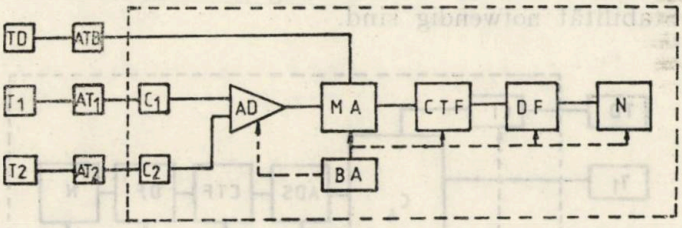

Abb. 2

A. Der Zähler hat al Eingangsgrößen vereinheitlichte elektrische Signale, proportional der Größen der Durchflußmengen und der Temperaturen des Hin- und Rückflusses (Abb. 2). In diesem Falle werden die primären Signale DT des Durchflusses und T_1 bzw. T_2 der Temperaturen, die man an den Ausgängen der Umwandler erhält, umgewandelt in vereinheitlichte Signale ATB für den Durchfluß und AT_1 bzw. AT_2 für die Temperaturen, was ausserhalb des Zählers durch elektronischen Adapter geschieht.

Die im Zähler vorhandenen elektronischen Stufen haben folgende Bedeutung : C_1, C_2 — Korrekturstufen für die Kompensation der Temperaturabhängigkeit der Durchflußparameter, AD — Differentialverstärker, MA — elektronischer Amplituden—Dauer-Multiplikator, CTF — Spannungs

— Frequenzkonvertor, *DF* — Frequenzteiler, *N* — Impulszähler und *BA* — gemeinsamer Speisespannungsblock.

Entsprechend der Variante A (Abb. 2) nimmt der Zähler vor der Produktbildung die Korrekturen vor, berechnet analog die Wärmemenge und zeigt das Ergebnis numerisch an. Obwohl dieses Schema sehr kostspielig ist, fügt es sich in die Typisierungsprinzipien von Baugruppen ein.

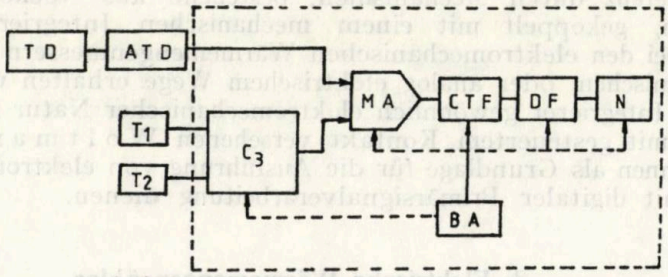

Abb. 3

B. Der Zähler hat als Eingangsgrößen primäre Signale für die Temperatur und vereinheitlichte Signale für die Dürchflußmenge (Abb. 3). In diesem Fall übernimmt der elektronische Block C_3 die Funktionen der Temperaturadapter und der Differentialverstärker und führt gleichzeitig auch die Korrektur durch. Die Ausführung dieser Zählerbaugruppe wirft komplexe Probleme auf, für deren Lösung Materialienteile erhörher Präzision und Stabilität notwendig sind.

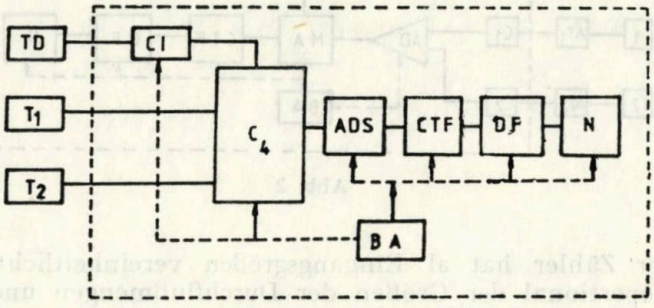

Abb. 4

C. Der Wärmemengenzähler benutzt als Eingangsgrößen primäre elektrische Signale (Abb. 4), die direkt von den Durchflußmengen und Temperaturwandlern kommen. In diesem Fall führt der elektronische Block C_4 die Multiplikations- und Korrekturoperationen aus, wobei der Eingangskreis *CI* den Ausgang des elektronischen Durchflußmengenumwandlers dem Eingang der multifunktionellen Stufe C_4 angepaßt wird.

Durch die erhöhte dynamische Abstufung des Mehrfunktionenblockes können die elektronischen Komponenten in Grenzsituationen arbeiten, was ein übertriebenes Anwachsen der Fehler am Anfang der Meßbereiche, sowohl für die Durchflußmengen als auch für die Temperaturen, zur Folge hat, eine Situation, die nur durch die Verwendung von integrierten Präzisionsverstärkern vermieden werden kann.

Das Prinzipschema aus Abb. 4 benutzend, kann man einen elektronischen Wärmemengenzähler erhalten, bei dem der Nichtlinearitätsfehler von 0,5% mit Thermowiderständen Cu 100 und von 1% mit Thermowiderständen Pt 100 nicht überschritten wird.

Unter den Bedingungen einer korrekten Eichung wird eine Zählergenauigkeit mit Umwandlern Pt 100 von 2,5% und bei Zählern mit den Umwandlern Cu 100 von 1,5% erreicht. Der Arbeitstemperaturbereich beträgt 50°...150°C für den Fluß auf der Hinleitung und 40°...100°C für die Rückflußleitung, wobei die maximale Temperaturdifferenz zwischen der Hin- und Rückflußleitung 100°C betragen kann.

Als Meßelement für die Warmwasserdurchflußmenge wurde ein elektromagnetischer Durchflußmengenumwandler vom Typ DEM—90, ein Erzeugnis von I.T.R.D., Paşcani, verwendet, der an die Hinleitung montiert wurde.

4. Schlußfolgerungen

a) Die Wärmemengenzähler für Warmwasserinstallationen integrieren über eine bestimmte Zeitspanne das Produkt von massemässiger Durchflußmenge und Temperaturdifferenz zwischen Hin- und Rückflußleitung in dynamischen Abstufungen von 1/10 für die Temperaturen und 1/20 für die Durchflußmengen.

b) Weil einige der Parameter der Flußmenge und die der Temperaturumwandler von der Temperatur abhängen, müssen die Präzisionszähler mit Korrekturstufen versehen sein, da die Nichtlinearitätsfehler bei kleinen Temperaturdifferenzen zwischen Hin- und Rückleitung groß sind.

c) Gegenüber anderen Typen von Durchflußmengenumwandlern kann der elektromagnetische Umwandler, da er eine erhöhte Meßgenauigkeit bei einem sehr breiten Durchmesserspektrum liefert als Grundlage dienen für die Konstruktion von elektronischen Wärmemengenzählern ohne bewegliche Elemente und kann unserer Meinung nach zu größtmöglichen Meßgenauigkeiten bei der Bestimmung der Wärmemenge führen.

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ASUPRA MĂSURĂRII CANTITĂŢII DE CĂLDURĂ TRANSPORTATĂ PRIN APĂ CALDĂ

(Rezumat)

Se analizează procedeele actuale de măsurare a cantităţii de căldură transportată printr-un fluid fără schimbare de fază şi, în pas cu dezvoltarea tehnicii construcţiei debitmetrelor, se propune o nouă clasificare a contoarelor de căldură. Sînt prezentate schemele bloc ale contoarelor electrice de căldură echipate cu debitmetre electromagnetice şi performanţele contorului realizat de către autori, cînd acesta lucrează cu traductoare Cu 100 sau Pt 100.

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C.Z. 621.314.6 ; 621.396.64

CIRCUITE REDRESOARE — AMPLIFICATOARE SENSIBILE LA FAZĂ REALIZATE CU AMPLIFICATOARE OPERAȚIONALE

DE

ADRIAN CREȚU

1. Introducere

Diversitatea schemelor de redresoare sensibile la fază (RSF) cu aplicații în tehnica măsurărilor electrice este relativ mare [1], ..., [4], [6]. Dintre acestea, însă, numai cele propuse de W e n g l e r [2] asigură simultan cu operația de redresare sensibilă la fază și amplificarea semnalului de măsurat.

Schemele studiate în [2], construite cu unul sau două tranzistoare bipolare, conectate în brațele unor punți cu rezistențe, au o serie de neajunsuri care le limitează domeniul de aplicare. Dintre acestea sînt de reținut următoarele : a) dependența puternică a semnalului de ieșire, redresat, atît de amplitudinea semnalului de măsurat cît și de cea a semnalului de comandă. Aceasta impune existența unui semnal de comandă cu amplitudinea riguros constantă, de preferință de formă rectangulară ; b) liniaritatea ecuației de funcționare se menține numai pentru variații în limite restrînse ale semnalului de măsurat și numai dacă rezistența de sarcină are valori foarte mari ; c) consumul de energie în circuitul de comandă este ridicat, comparabil cu cel al RSF construite cu diode semiconductoare etc.

În cele ce urmează se propune și se studiază schema de principiu a unui redresor—amplificator sensibil la fază (RASF), avînd ca element de bază un amplificator operațional (AO). Se propun, de asemenea, două scheme practice de RASF cu AO, una cu bornă comună între circuitele de comandă și de măsurare și alta avînd cele două circuite „decuplate electric", care răspund satisfăcător majorității aplicațiilor RSF din tehnica măsurătorilor.

2. Schema electrică de principiu a RASF. Ecuația de funcționare

Schema electrică de principiu a RASF este reprezentată în fig. 1.

Circuitul de comandă al RASF cuprinde în esență două contacte, K_1 — normal închis și K_2 — normal deschis, a căror stare este determinată de polaritatea semnalului de comandă, $u_c(\omega t)$, care are aceeași frecvență cu semnalul de măsurat, $u_M(\omega t)$.

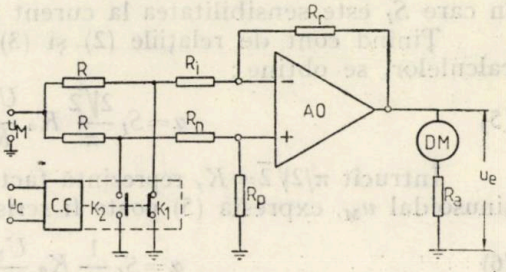

Fig. 1

Semnalul de măsurat , u_M, este aplicat, în funcție de starea contactelor K_1 și K_2, la una dintre intrările amplificatorului operațional diferențial cu reacție negativă și cu amplificări egalizate, AO. Astfel, în alternanța semnalului u_c în care contactul K_1 este închis iar contactul K_2, deschis, semnalul u_M este aplicat numai la intrarea neinversoare „+" a AO, intrarea „—" a AO fiind conectată în acest caz la masă (prin K_1). Analog, în cealaltă alternanță a semnalului u_c, în care poziția celor două contacte se inversează, același semnal u_M este comutat la intrarea inversoare „—" a AO. Pentru ca amplificarea AO să rămînă practic aceeași, indiferent de poziția contactelor K_1 și K_2, trebuie îndeplinită condiția [5] :

$$\frac{R_r}{R+R_i} = \frac{R_p}{R+R_n} = K_a. \tag{1}$$

Fie :

$$u_c = \sqrt{2}\, U_c \sin \omega t, \quad u_M = \sqrt{2}\, U_M \sin(\omega t - \varphi), \tag{2}$$

expresiile analitice ale semnalelor de comandă, respectiv de măsurat ale RASF, defazate între ele cu un unghi φ, reglabil în limitele [0, 180°].

În ipoteza, ușor de realizat practic, că amplificările pe buclă ale AO sînt mult mai mari decît unitatea, valoarea instantanee a semnalului de ieșire, $u_e(\omega t)$, al RASF, va rezulta din [5] :

$$\begin{cases} u_e(\omega t) = K_a u_M(\omega t) & \text{— în intervalul } \omega t \in (0, \pi), \text{ în care semnalul } u_c > 0, \\ & (K_1 \text{ — închis ; } K_2 \text{ — deschis) ;} \\ u_e(\omega t) = -K_a u_M(\omega t) & \text{— în intervalul } \omega t \in (\pi, 2\pi), \text{ în care semnalul} \\ & u_c < 0, \ (K_1 \text{ — deschis ; } K_2 \text{ — închis).} \end{cases} \tag{3}$$

Coeficientul de amplificare, K_a, al AO diferențial cu reacție negativă și cu amplificări egalizate, utilizat în acest caz, poate fi atît unitar cît și supraunitar, în funcție de aplicațiile preconizate pentru RASF, valoarea sa stabilindu-se pe baza condiției (1).

Deviația α a dispozitivului magnetoelectric DM (microampermetrul μA), înseriat împreună cu rezistența adițională R_a în circuitul de ieșire al RASF, proporțională cu valoarea medie a semnalului u_e/R_a, calculată pe o perioadă a semnalului de comandă u_c, va rezulta din :

$$\alpha = S_1 \frac{U_{c\,\mathrm{med}}}{R_a} = S_I \frac{1}{R_a} \cdot \frac{1}{2\pi} \int_0^{2\pi} u_e(\omega t)\, \mathrm{d}\omega(t), \tag{4}$$

în care S_I este sensibilitatea la curent a dispozitivului DM.

Ținînd cont de relațiile (2) și (3), din (4), după efectuarea tuturor calculelor, se obține :

$$\alpha = S_I \frac{2\sqrt{2}}{\pi} K_a \frac{U_M}{R_a} \cos \varphi. \tag{5}$$

Întrucît $\pi/2\sqrt{2} = K_f$ reprezintă factorul de formă al curbei semnalului sinusoidal u_M, expresia (5) poate fi scrisă sub forma :

$$\alpha = S_I \frac{1}{K_f} K_a \frac{U_M}{R_a} \cos \varphi. \tag{6}$$

Relația (6) diferă de ecuația de funcționare generală a RSF cu redresare dublă, dată de [3] :

$$\alpha = S_I \frac{1}{K_f} \cdot \frac{U_M}{R_a} \cos \varphi, \tag{7}$$

numai prin faptul că sensibilitatea RASF propus este multiplicată cu coeficientul $K_a \geqslant 1$, rezultat din (1).

Ecuația de funcționare (6) arată că RASF, reprezentat în fig. 1, poate fi utilizat în majoritatea aplicațiilor binecunoscute ale RSF [3], avînd în raport cu acestea avantajul amplificării directe a semnalului de măsurat, u_M.

3. Realizări practice

Se propun, în continuare, două scheme practice prin care se poate materializa schema de principiu din fig. 1.

Prima schemă (fig. 2) constituie un RASF cu bornă comună (masa) între sursele semnalelor u_M și u_c, contactele K_1 și K_2 fiind realizate cu ajutorul a două tranzistoare complementare, T_1 — de tip *npn*, respectiv T_2 — de tip *pnp*, care funcționează în regim de comutație.

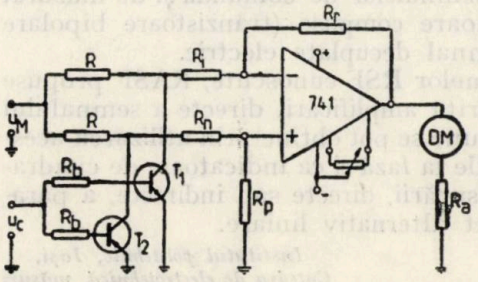

Fig. 2

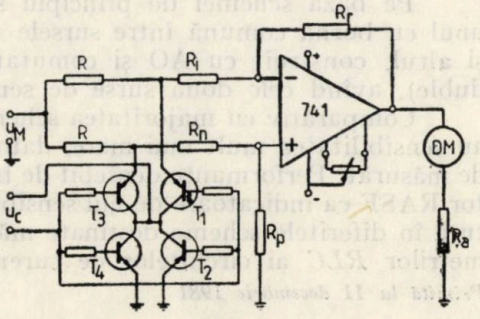

Fig. 3

Principalul neajuns al acestei scheme constă în faptul că, în scopul veitării regimului saturat inversat al celor două tranzistoare, este necesară îndeplinirea condiției restrictive :

$$U_c \gg U_M. \tag{8}$$

Cea de a doua schemă (fig. 3) înlătură dezavantajul menționat, întrucît contactele K_1 și K_2 sînt realizate cu ajutorul a două comutatoare compuse constituite prin conectarea serie—opoziție a perechilor de tranzistoare complementare T_1, T_2 — de tip *pnp*, respectiv T_3, T_4 — de tip *npn*.

Schema RASF din fig. 3, ca de altfel marea majoritate a RSF cunoscute, are sursele semnalelor de măsurat și de comandă decuplate electric.

Tipurile AO și tranzistoarelor $T_1, \ldots, T_4$, precum și valorile rezistențelor din schemele RASF se vor stabili în raport cu destinația acestora. RASF propuse în prezenta lucrare, utilizate în principal ca indicatoare de nul sensibile la fază și ca detectoare de cuadratură în diverse scheme destinate măsurării parametrilor *RLC* ai circuitelor de curent alternativ liniare, sînt construite cu amplificatoare operaționale de tipul μA741 și tranzistoare

bipolare complementare, cu siliciu, de tipul BC107, respectiv BC177. Pentru RASF din fig. 3 este evident avantajoasă utilizarea tranzistoarelor bipolare duble de tip *pnp*, respectiv *npn*.

Valorile rezistențelor R_i, R_n, R_r și R_p din schemele RASF se stabilesc pe baza condiției (1), astfel încît deriva de zero a RASF, corespunzătoare situației în care $u_M=0$ și $u_c\neq 0$, să fie practic nulă. În acest scop, pentru comoditatea anulării derivei, rezistențele R_i și R_n pot fi semireglabile. Întrucît, indiferent de starea contactelor K_1, K_2, impedanța de intrare a RASF va fi practic egală cu rezistența R, valoarea celor două rezistențe se adoptă în funcție de destinația RASF.

Principalele surse de erori ale RASF propuse sînt determinate de regimurile tranzitorii de comutație ale contactelor K_1 și K_2. Aceste erori, mai mari în cazul RASF din fig. 2 și maxime la $\varphi=\pm\pi/2$, devin neglijabile dacă $U_c \gg U_M$ sau dacă semnalul de comandă u_c este dreptunghiular.

4. Concluzii

Se propune și se studiază, stabilindu-se ecuația de funcționare, schema de principiu a unui RASF avînd ca element de bază un amplificator operațional cu reacție negativă și cu amplificări egalizate.

Pe baza schemei de principiu se propun două RASF, dintre care unul cu bornă comună între sursele semnalelor de comandă și de măsurat și altul, construit cu AO și comutatoare compuse (tranzistoare bipolare duble), avînd cele două surse de semnal decuplate electric.

Comparativ cu majoritatea schemelor RSF cunoscute, RASF propuse au sensibilitatea mult mai mare, datorită amplificării directe a semnalului de măsurat. Performanțe deosebit de bune se pot obține prin utilizarea acestor RASF ca indicatoare de nul sensibile la fază și ca indicatoare de cuadratură în diferitele scheme destinate măsurării, directe sau indirecte, a parametrilor *RLC* ai circuitelor de curent alternativ liniare.

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Institutul politehnic, Iași,
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CIRCUITS REDRESSEURS—AMPLIFICATEURS SENSIBLES À LA PHASE, RÉALISÉS AVEC DES AMPLIFICATEURS OPÉRATIONNELS

(Résumé)

On propose et on étudie le schéma de principe et deux schémas pratiques de quelques redresseurs — amplificateurs sensibles à la phase, construits à réaction negative et ayant les facteurs d'amplification égalisés. Ces dispositifs peuvent être utilisés, avec de bonnes performances, comme indicateurs de zéro sensibles à la phase et comme détecteurs de quadrature dans les schémas de mesure pour les paramètres R, L, C des circuits linéaires de c.a.

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D.K. 621.317.44

DIE ANTWORTSKURVE EINES VERSCHIEBUNGSGEBERS MIT MAGNETISCHER SONDE, MIT SENKRECHTER BEWEGUNG DES GLEICHMAGNETS ZUR SONDENACHSE

VON

PARASCHIVA HABA

1. Einführung

In einer früheren Arbeit [1] wurde die Möglichkeit der Konstruktion eines Verschiebungsgebers untersucht. In [2] wurde ein Verschiebungsgeber mit magnetischer Sonde mit entlanger Bewegung des Gleichmagnets zur Sondenachse einer theoretischen Analyse unterworfen. Dabei wurde folgender Zusammenhang zwischen dem gemessenen Strom I und dem Steuerungsfluß Φ_c aufgestellt:

$$I = \frac{U}{2R\Phi_m}\,\Phi_c, \tag{1}$$

wobei: U — die mittlere Betriebsspannung des Meßkreises, R — den elektrischen Wiederstand eines Zweiges des Meßkreises, Φ_m — den maximalen Wert des Flusses im Durchschnittsfläche eines Zweiges der Sonde bedeuten.

In der erwähnten Arbeit wurde der Steuerungsfluß als Funktion des Abstandes x zwischen der Sonde und dem Magnet mittels der zwischen den Polen des Gleichmagnets wirkenden magnetischen Gesamtleitfähigkeit des Luftraumes G_m ausgedrückt.

2. Theoretische Betrachtungen

In dieser Arbeit wollen wir ein Verschiebungsgeber mit magnetischer Sonde untersuchen, bei dem der Gleichmagnet nicht entlang, sondern senkrecht zur Sondenachse verschoben wird. Für diesen Geber wollen wir den Zusammenhang zwischen dem gemessenen Strom I und der Magnetverschiebung x auf eine andere Weise wie in [2] ableiten.

Im Prinzip sieht ein solcher Verschiebungsgeber wie in der Abb. 1 aus. In diesem Falle besteht der Steuerungsfluß aus einem kleinen Teil des Haupt- und Randflußes. Der restliche Fluß beeinflußt die Funktionierung der Sonde nicht.

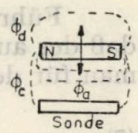

Abb. 1

Als maximaler Steuerungsfluß wird derjenige betrachtet, der durch die Sonde bei den kleinsten Abstand zwischen der Sonde und dem Magnet durchgeht.

Für die in den Abb. 1 angegebene Verteilung der Flußlinien, die einem gegebenen Abstand zwischen der Sonde und dem Magnet entspricht, darf man eine äquivalente Schaltung (Abb. 2) angeben, die aus einer Reihenkette von Vierpolen besteht.

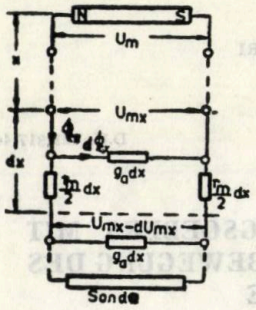

Abb. 2

In dieser Schaltung bedeuten : g_a — die magnetische Leitfähigkeit pro Längeeinheit des Streuflusses durch die Luft, r_m — der magnetische Wiederstand pro Längeeinheit des durch die Sonde gehenden magnetischen Flußes.

Für Vereinfachung wird angenommen, daß sich r_m und g_a entlang der Reihenkette nicht verändern.

Betrachtet man nun ein Element des Ersatzbildes von der Länge dx beim Abstand x, so kann man schreiben :

$$(1)\quad -U_{mx}+r_m\,dx\,\Phi_{cx}-r_m\,dx\,d\Phi_{cx}+(U_{mx}-dU_{mx})=0.$$

Vernachlässigt man weiter die Glieder zweiter Ordnung, so bekommt man :

$$(2)\quad \frac{dU_{mx}}{dx}=r_m\,\Phi_{cx}.$$

Da aber $d\Phi_{cx}=U_{mx}g_a\,dx$ ist, so erhält man :

$$(3)\quad \frac{d\Phi_{cx}}{dx}=g_aU_{mx}.$$

Aus den Beziehungen (2) und (3) ergibt sich am Ende :

$$(4)\quad \frac{d^2\Phi_{cx}}{dx^2}-k^2\Phi_{cx}=0,$$

worin $k^2=r_mg_a$ eine Konstante ist.

Die Lösung dieser Differentialgleichung ist allgemein bekannt und lautet :

$$(5)\quad \Phi_{cx}=C_1e^{kx}+C_2e^{-kx},$$

wobei sich die Konstanten C_1 und C_2 aus den folgenden Bedingungen ergeben :

$$\Phi_c(0)=\Phi_{cmax}\text{ und }\Phi_c(\infty)=0.$$

Führt man diese Bedingungen in (5) ein, so erhält man für Φ_{cx} den folgenden Ausdruck :

$$(6)\quad \Phi_{cx}=\Phi_{cmax}\,e^{-kx}.$$

Führt man weiter (6) in (1) ein und hält die Tatsache vor Augen, daß der auf eine Zweige wirkende Steuerungsfluß gleich $\Phi_{cx}/2$ ist, so erhält man für den zu messenden Strom den Ausdruck :

$$(7)\quad I=\frac{U}{4R\Phi_m}\Phi_{cmax}\,e^{-kx}=I_{max}\,e^{-kx},$$

wobei $I_{max}=\frac{U}{4R}\cdot\frac{\Phi_{cmax}}{\Phi_m}$, den maximallen Strom darstellt.

3. Experimentelle Ergebnisse

Gemäß des obengeschilderten Prinzips wurde ein Verschiebungsgeber aufgebaut, dessen Ersatzbild in Abb. 3 wiedergegeben ist. Die Sonde ist mit zwei Wicklungen je 2 000 Windungen vorgesehen und der Gleichmagnet ist der Länge 50 mm und des Durchmessers 10 mm. Die Bewegung des Magnets gegenüber der Sonde wird auf eine milimetersche Linie verfolgt. Wegen der Dicke der Wicklungen kann der Magnet nur bis 5 mm Abstand zu der Sonde angenähert werden.

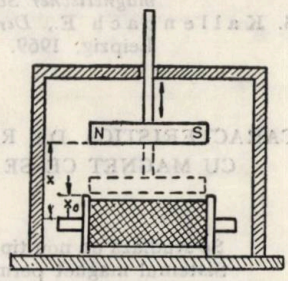

Abb. 3

Es wurden mehrere Antwortskurven der Sonde, $I=f(x)$, aufgenommen und untereinander gemittelt. Die gemittelte Antwortskurve ist graphisch in der Abb. 4 angegeben. Um zu sehen, inwieweit diese Kurve dem Ausdruck (7) entspricht, haben wir $\ln I$ gegen x in der Abb. 5 wiederum aufgetragen. Wie man Sieht, ist diese Kurve eine Gerade, deren Neigung und Schnitt die Werte haben $-0{,}07$ mm^{-1} bzw 3,05. Hieraus ergibt sich für den gemessenen Strom der Ausdruck;

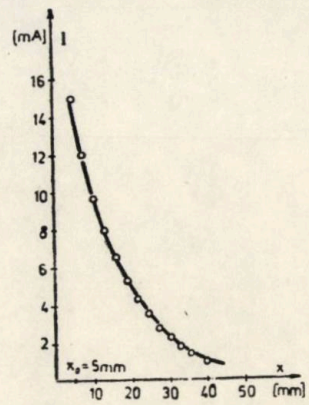

Abb. 4

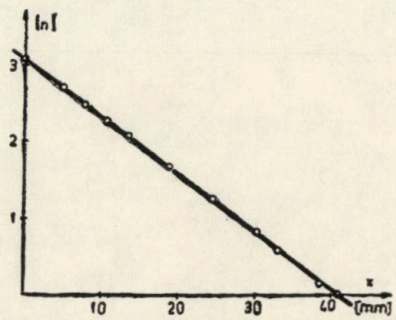

Abb. 5

$$I=21{,}6\ e^{-0{,}07x}. \tag{8}$$

Ähnliche experimentelle Beziehungen wurden auch für andere Verschiebungsgeber dieser Art gefunden.

Zusammenfassend darf man sagen, daß die in dieser Arbeit angenommenen Hypothesen von den experimentellen Daten bestätigt sind.

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Technische Hochschule, Jassy, Lehrshl fur Elektrotechnik, elektrische Messungen und Maschinen

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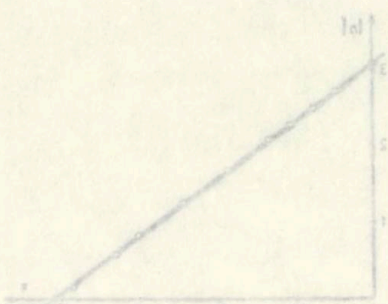
CARACTERISTICA DE RĂSPUNS A UNUI TRADUCTOR CU SONDA MAGNETICĂ CU MAGNET CE SE DEPLASEAZĂ PERPENDICULAR PE AXUL SONDEI

(Rezumat)

Se studiază un nou tip de traductor cu sondă magnetică comandată cu magnet permanent.

Sistemul magnet permanent — sondă se reprezintă într-o schemă echivalentă constituită dintr-un şir de cuadripoli (fig. 2). Se deduce expresia fluxului de comandă în funcţie de distanţa x, de formă exponenţială. Se înlocuieşte în expresia curentului (1) şi se găseşte pentru acesta expresia (7).

Cu un dispozitiv realizat experimental se trasează caracteristica de răspuns $I=f(x)$. Cu aceste date se calculează şi se trasează $\ln I=f(x)$ (fig. 5) şi se găseşte că forma acestei caracteristici este liniară. De aici se trage concluzia că rezultatele experimentale confirmă ipotezele teoretice.



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C.D. 621.317.336

APPAREIL DE MESURE DE LA SALINITÉ DES SOLS

PAR

TRAIAN BALAN, DANA BALAN et FLOREL MATEI

1. Introduction

L'exploitation et le traitement adéquat des sols salines impliquent la connaissance précise de la concentration et de l'étendue de ces sols.

Les observations visuelles des sols et des plantes cultivées, en général, ne permettent pas d'établir le diagnostic de la salinité parce que la concentration des sels peut réduire la production des cultures jusqu'à 25% sans qu'il existe des signes indicatifs apparents.

Actuellement le diagnostic correct du degré de salinité implique l'analyse des échantillons de sol apportés dans le laboratoire. L'utilisation de la trousse portative d'analyse réduit le temps de travail mais l'effort reste remarquable dû au nombre élevé d'échantillons nécessaires et au temps nécessaire pour l'analyse.

Pour estimer l'efficience des méthodes d'exploitation on recommande le contrôle périodique du degré de salinité, ce dernier variant avec l'administration de certains types d'engrais.

Durant la dernière décennie ont été mises au point diverses méthodes de détermination du degré de salinité basées sur les méthodes électriques de détermination de la résistivité des sols [1]...[3]. Même si les méthodes ne précisent pas le type de salinité on impose, par commodité, aussi, à préciser les déterminations sur les terrains avec un certain type de salinité ou dans la surveillance du degré de salinité en temps (dans les serres).

2. Mesure de la résistivité des sols

La conductivité électrique des sols salines est généralement de nature électrolytique, par l'intermède de l'eau des interstices contenant des électrolytes dissouts, et croît, vu les cations changeants et le Na soluble existant dans les sols argileux. Ceux-ci conduisent à une conductivité élevée des couches superficielles des sols. Pour un type donné de sol on peut déterminer un lien entre sa salinité et sa conductivité, surtout si le contenu en eau est relativement constant (sols irrigués, serres etc.)

Le schéma de mesure le plus utilisé pour la détermination de la résistivité des sols est une variante de la méthode de compensation avec quatre electrodes — le schéma B e h e r e n d, méthode de W e n n e r. Celle-ci

présente l'avantage d'une précision et d'une sensibilité élevée, sans danger d'électrocution et d'une consommation réduite.

Le schéma bloc de l'appareil est représenté dans la Fig. 1. En principe on envoie par la résistance R_p un c.a., par des électrodes de courant A et B et on compare la chute de tension U_p avec celle aux bornes d'une résistance

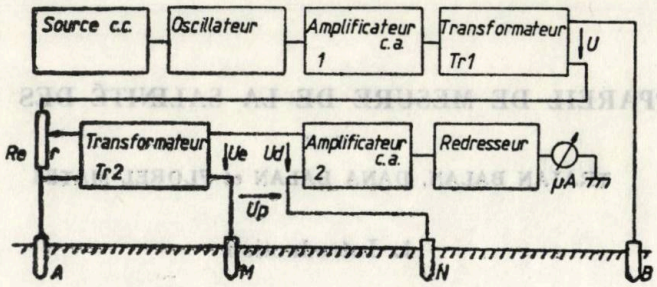

Fig. 1

connue R_e parcourue par le même courant I_p. L'égalité des deux tensions est vérifiée grâce à un indicateur de nul connecté dans le circuit à l'aide d'un électrode de potentiel.

2.1. Description de l'appareil

L'appareil (Fig. 2), contient les blocs ci-dessous :

1° *Bloc d'alimentation,* qui contient :

a) Source de c.c. de $\pm$ 9 V, obtenue à l'aide de 4 piles 3R-12.

b) Oscillateur — convertisseur c.c. — c.a. Par le schéma réalisé (circuit operationnel βA-741, un circuit *RC* oscillant dans le réseau W i e n), l'oscillateur — convertisseur assure un volume et une consommation sen-

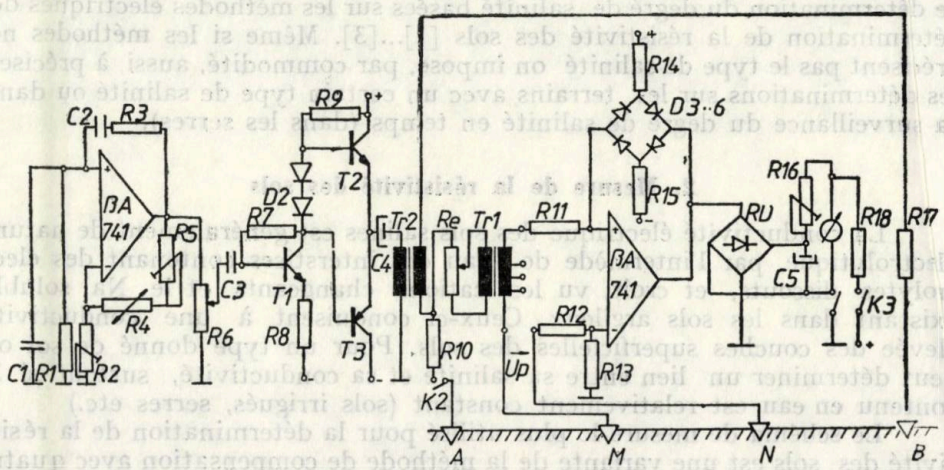

Fig. 2

siblement réduite par rapport à celles réalisés avec des éléments discrets, une grande stabilité de la fréquence et de l'amplitude de la tension pratiquement sinusoïdale. La fréquence de l'oscillateur a été fixée à 72 ± 1 Hz pour réduire l'influence des courants alternatifs parasites du sol (le sol se comporte comme un filtre passe-bas). Il n'est pas recommandable un accroissement de la fréquence au-delà de 100 Hz, la construction en nécessitant des noyaux magnétiques spéciaux.

c) Amplificateur de puissance (classe B) en contre-temps, pour l'alimentation du primaire du transformateur séparateur. L'amplificateur est réalisé avec des transistors complémentaires BD-137, BD-138.

d) Transformateur de séparation (2,5 V/20 V) qui assure l'obtention d'un signal suffisamment puissant aux mesures.

2° *Bloc de mesure*, constitué par un transformateur de mesure et d'un amplificateur différentiel avec indicateur de nul.

a) Le transformateur de mesure a le rôle de créer un déphasage de la chute de tension de compensation du potentiomètre R_e de 180°. Dans le cas où la résistance R_e est un potentiomètre bobiné multitour, étalonné (10 kΩ, 10 tours), les prises du transformateur (30/1, 10/1, 3/1, 2/1) assurent la mesure de la résistance R_p en quatre domaines de mesure (0,1...1 Ω ; 1...10 Ω ; 10...100 Ω ; 100...500 Ω). La valeur de la résistance R_p des sols varie de 0,3 Ω (pour les sols trés salés) jusu'à 400 Ω (le sable sec). En général, la résistance R_p est comprise entre 1...40 Ω, ce qui corespond à une résistivité $\rho = 10...400$ Ωcm.

b) L'amplificateur différentiel avec redresseur et appareil indicateur de nul (100 μA) permettent à constater l'équilibre de deux tensions (inconnue et étalon). L'amplificateur présente la particularité de croître son amplification à l'entrée des faibles signaux. Ceci s'opère grâce au pont des diodes $D_1...D_4$, en anneau, avec une diagonale apliquée entre la sortie de l'amplificateur et l'entrée non-invertrice, l'autre diagonale étant connectée à la source d'alimentation.

Comme système de redresseur on peut utiliser un pont Graetz ou au choix un amplificateur et un redresseur double-alternance, (Fig. 3), très efficace dans le cas des faibles signaux.

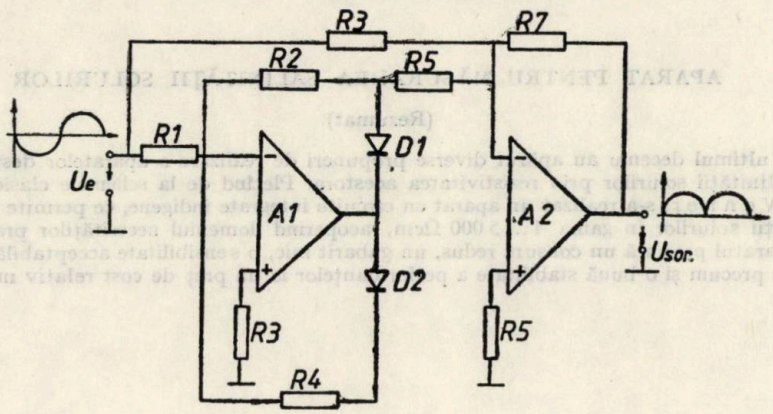

Fig. 3

L'appareil a la possibilité de fonctionner jusqu'à une tension d'alimentation de 80% U_n.

Après six mois d'utilisation l'appareil présente une fiabilité élevée, une précision d'environ 1% et dans les conditions d'un étalonnage initial correct, une sensibilité satisfaisante.

L'appareil peut aussi servir de ohmmètre dans la bande 0,1...500 Ω par le courtcircuit des bornes $A-M$ et $B-N$.

Par l'utilisation de certains électrodes de dimensions différentes [2] on peut établir le profil de la salinité des sols en profondeur.

3. Conclusions

1. L'appareil réalisé est recommandé aux entreprises agricoles aussi qu'aux recherches de specialité.

2. L'appareil permet la mesure de la résistivité des sols dans la gamme de 0,1...500 Ω, couvrant le domaine des nécessités pratiques.

3. Il est réalisé avec des composants intégrés indigènes, ce qui lui assure une consommation réduite, un petit gabarit, une bonne stabilité des performances et un prix de vente acceptable.

4. Avec un étalonnage initial correct, les résultats peuvent se réproduire assurant la sensibilité imposée par l'utilisateur.

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APARAT PENTRU MĂSURAREA SALINITĂŢII SOLURILOR

(Rezumat)

În ultimul deceniu au apărut diverse propuneri de realizare a aparatelor destinate măsurării salinităţii solurilor prin rezistivitatea acestora. Plecînd de la schemele clasice Behe r e n d—W e n n e r, s-a realizat un aparat cu circuite integrate indigene, ce permite măsurarea rezistivităţii solurilor în gama 1 ... 5 000 Ωcm, acoperind domeniul necesităţilor practice.

Aparatul prezintă un consum redus, un gabarit mic, o sensibilitate acceptabilă (ce poate fi extinsă) precum şi o bună stabilitate a performanţelor la un preţ de cost relativ mic.

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ON THE POSSIBILITY OF DIGITAL COMPUTING BLOCKS SYNTHESIS BY USING THE IZOMORFISM IN CODING

BY

VALERIU MUNTEANU, GH. ZAHARIA and LIVIU GORAŞ

1. Introduction

It is often necessary to implement digital electronic circuits able to perform different mathematical operations (addition, substraction, multiplication, raising to powers etc.) between elements belonging to a finite number p of sets X_k, $(k=\overline{1,p})$.

These operations can be internal or external composition laws. As any digital computing block has a finite number of states, the sets X_k, $(k=\overline{1,p})$, have a finite number of elements, that is

$$\text{card } X_k < \infty, \ (k=\overline{1,p}). \tag{1}$$

The internal operations (composition laws) will be denoted by

$$\pi_k : X_k \times X_k \to X_k, \ (k=\overline{1,p}) \tag{2}$$

and the external ones by

$$\varepsilon_{rs} : X_r \times X_s \to X_s, \ (r \neq s, \ r=\overline{1,p}, \ s=\overline{1,p}). \tag{3}$$

The outputs of the electronic digital circuits will have a finite number of logical states that will be generally denoted by the set S of the form

$$S=\{s_1, s_2, \ldots, s_m\}, \tag{4}$$

where

$$m=\text{card } S < \infty. \tag{5}$$

In order to be able to perform mathematical operations using digital electronic circuits it is necessary to establish a code i.e. a mapping between elements belonging to X_k and elements from S. This coding is done using certain coding functions f_k. If we denote by $f_k(x_i)$ the coded word corresponding to the element $x_i \in X_k$, between the coded words we have the operations Π_k, Σ_{rs}, operations that must „correspond" to π_k, respectively ε_{rs} given by (2) and (3).

2. The Izomorphism in Coding

Let X_k be a set having a finite number of elements, that is

$$n = \text{card } X_k < \infty, \tag{6}$$

so that this set can be represented in the form

$$X_k = \{x_1, x_2, \ldots, x_n\}. \tag{7}$$

To any $x_i \in X_k$, $(i = \overline{1, n})$, must correspond a code word, c_i, having a length of l_i, the symbols of which the code words are made belong to the logical state set S, given by (4). Of course the mapping between elements $x_i \in X_k$ and c_i must be one to one [1],...,[3].

Considering the codes as having the length l, we have

$$l_i = l, \quad (i = \overline{1, n}). \tag{8}$$

In this way the code is instantaneous too [2], [3]. The necessary and sufficient condition for the existence of instantaneous codes is given by K r a f t's inequality [2], [3]:

$$\sum_{i=1}^{n} m^{-l_i} \leqslant 1. \tag{9}$$

Taking into consideration (8), the relation (9) becomes

$$n m^{-l_i} \leqslant 1 \tag{10}$$

or

$$\log_m n \leqslant l. \tag{11}$$

The length of the code words can be evaluated with the relation

$$l = -[-\log_m n], \tag{12}$$

where $[A]$ represents the integer part of the real number A.

Let S^l be the l order extension of the set S [3] and the function f_k defined as

$$f_k : X_k \to S^l. \tag{13}$$

The code being instantaneous, that is uniquely decodable, the function f_k, $(k = \overline{1, p})$, must be a bijection.

From

$$n \leqslant m^l \tag{14}$$

we have

$$\text{card } X_k \leqslant \text{card } S^l. \tag{15}$$

If

$$n < m^l \tag{16}$$

so that

$$\text{card } X_k < \text{card } S^l, \tag{17}$$

the function f_k is not a surjection.

In this situation the coding will be done be done by means of the function

$$\tilde{f}_k : X_k \to f_k(X_k) \subset S^l, \tag{18}$$

defined by

$$\tilde{f}_k(x_i)=f_k(x_i),\ \forall x_i\in X_k,\ (i=\overline{1,n}\ ;\ k=\overline{1,p}). \tag{19}$$

In this way $\tilde{f}_k$ is a surjection. Taking into consideration the fact that f_k is biunivocal, by virtue of (19) we have that $\tilde{f}_k$ is a bijection.

The most common case used in digital computing corresponds to

$$m=2. \tag{20}$$

In this case the set S has two elements, s_1 and s_2. To the element s_1 corresponds a low voltage (current) level denoted by L (low) or logical zero and to the element s_2 a high voltage (current) denoted by H (high) or logical one.

For $m=2$ we shall denote the set S by

$$S=\mathcal{L}=\{L, H\}. \tag{21}$$

In the following we shall consider only this case. The number of biunivocal functions f_k or bijective $\tilde{f}_k$ can be determinated by means of the relation

$$N_f=A_m^n l, \tag{22}$$

where A_k^i denotes arrangements of k elements taken by i, with $k, i\in N$, $i\leqslant k$.

Different digital electronic circuits can be obtained corresponding to different functions f_k among the N_f possible functions. The equivalence between the operation π_k and Π_k, respectively ε_{rs} and Σ_{rs}, are in fact the izomorphism conditions :

$$\tilde{f}_k \text{ — bijections,} \tag{23}$$

$$\tilde{f}_k(x\pi_k y)=\tilde{f}_k(x)\Pi_k\tilde{f}_k(y),\ \forall x, y\in X_k,\ (k=\overline{1,p}), \tag{24}$$

$$\tilde{f}_r(x\varepsilon_{rs}y)=f_r(x)\ \Sigma_{rs}\tilde{f}_s(y),\quad \forall x\in X_r,\ \forall y\in X_s,\ (r\neq s, r=\overline{1,p}\ s=\overline{1,p}). \tag{25}$$

Relation (23) is satisfied by means of establishing the coding function that assures a code uniquely decodable.

Relations (24) and (25) permit the determination of the operations Π_k and Σ_{rs} that must be performed by the digital electronic circuit.

3. Implementation of the Digital Electronic Circuits

Supposing that the set X_k contains n elements, then, by means of relation (12), can be calculated the integer l that specifies the length of the code words. The l order extension of the set $\mathcal{L}$ (21) is a set containing 2^l elements that will be denoted by σ_j, $(j=1, 2, .., 2^l)$. Each element σ_j, $(j=1, 2, ..., 2^l)$, is formed of a succesion of l symbols L and/or H :

$$\sigma_1=LL\ ...\ LL,\ \sigma_2=LL...LH,\ \sigma_3=LL...HL,...,\ \sigma_{2^l}=HH...HH. \tag{26}$$

If $n<2^l$ we shall chose a coding function $\tilde{f}_k$ that establishes a bijective correspondence between the elements $x_i\in X_k$ and n arbitrary elements σ_j, $(j=1, 2,..., 2^l)$.

Being given the operation π_k that has to be implemented with a digital electronic circuit we give all the values for the variables $x \in X_k$ and, by means of the coding funtion $\tilde{f}_k$ that has been adopted, we establish the relations through the operation Π_k between the elements σ_j, $(j=1, 2,... 2^l)$, in concordance with (24).

The digital electronic circuit will have $2l$ input variables that will be denoted by $x_1, x_2,..., x_l, y_1, y_2, ..., y_l$ and l output variables denoted by $z_1, z_2, ..., z_l$.

For every output variable z_i, $(i=\overline{1, l})$, a Veitch-Karnaugh diagram [4] is made taking into account the relations obtained through the operation Π_k. In the case $n<2^l$ a number of in different states

$$N_k=2^{2l}-n^2, \tag{27}$$

will occur.

Exemple

Let us consider the set X_k of the form

$$X_k=\{-1, 0, 1\}$$

on which we define the internal operation π_k that coincide with the classical product in the set Z. In this case the operation π_k is defined in Table 1.

Table 1

π_k	-1	0	1
-1	1	0	-1
0	0	0	0
1	-1	0	1

In our exemple $n=$ card $X_k=3$ and by virtue of (12) for $m=2$ it results that the length of the code words is equal to

$$l=-[-\log_2 3]=-[-1{,}585\ldots]=2.$$

The second order extension of the set $\mathcal{L}$ is then

$$\mathcal{L}^2=\{LL, LH, HL, HH\}.$$

We adopt a coding function

$$\tilde{f}_k=\begin{pmatrix} -1 & 0 & 1 \\ \downarrow & \downarrow & \downarrow \\ HL & LL & LH \end{pmatrix}.$$

Relations (21) lead to the following system

$$\tilde{f}_k((-1)(-1))=\tilde{f}_k(-1)\Pi_k\tilde{f}_k(-1),\ \tilde{f}_k((-1)(0))=\tilde{f}_k(-1)\Pi_k\tilde{f}_k(0),\ \tilde{f}_k((-1)(1))=\tilde{f}_k(-1)\Pi_k\tilde{f}_k(1),$$

$$\tilde{f}_k((0)(-1))=\tilde{f}_k(0)\Pi_k\tilde{f}_k(-1),\ \tilde{f}_k(0, 0)=\tilde{f}_k(0)\Pi_k\tilde{f}_k(0),\ \tilde{f}_k(0, 1)=\tilde{f}_k(0)\Pi_k\tilde{f}_k(1),$$

$$\tilde{f}_k(1, (-1))=\tilde{f}_k(1)\Pi_k\tilde{f}_k(-1),\ \tilde{f}_k(1, 0)=\tilde{f}_k(1)\Pi_k\tilde{f}_k(0),\ \tilde{f}_k(1, 1)=\tilde{f}_k(1)\Pi_k\tilde{f}_k(1).$$

Taking into account the coding function adopted we obtain Table 2.

Table 2

z_1	z_2		x_1	x_2		y_1	y_2
L	H	$=$	H	L	Π_k	H	L
L	L	$=$	H	L	Π_k	L	L
H	L	$=$	H	L	Π_k	L	H
L	L	$=$	L	L	Π_k	H	L
L	L	$=$	L	L	Π_k	L	L
L	L	$=$	L	L	Π_k	L	H
H	L	$=$	L	H	Π_k	H	L
L	L	$=$	L	H	Π_k	L	L
L	H	$=$	L	H	Π_k	L	H

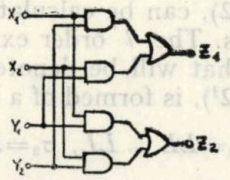

Fig. 1

The Veitch-Karnaugh diagrams for the output variables z_1 and z_2 are given in Table 3 and Table 4.

Table 3

x_1x_2 \ y_1y_2	LL	LH	HH	HL
LL				Ø
LH			Ø	H
HH	Ø	Ø	Ø	Ø
HL		H	Ø	

$z_1 = x_1y_2 \cup x_2y_1$

Table 4

x_1x_2 \ y_1y_2	LL	LH	HH	HL
LL				Ø
LH		H	Ø	
HH	Ø	Ø	Ø	Ø
HL			Ø	H

$z_2 = x_2y_2 \cup x_1y_1$

In Tables 3 and 4 the indifferent states have been doted by Ø and the logical operation „or" by $\cup$. The logic electronic circuit is represented in Fig. 1. In order to determine the external operation Σ_{rs} we assume a set X_r with a finite number n of elements and the set X_r with a finite number q of elements with $r \neq s$.

By means of (12) the length of the code words l_r for the set X_r and the length l_s of the code words for the set X_s can be calculated ($m=2$). If $n < 2^{l_r}$ and $q < 2^{l_s}$ we adopt the coding functions $\tilde{f}_r$ and $\tilde{f}_s$ that establishes a bijection between $x_i \in X_r$, $(i = \overline{1, n}\,;\ r = \overline{1, p})$, and n arbitrary elements of the l_r order extension of the set $\mathcal{L}$, respectively between the elements $y_j \in X_s$, $(j = \overline{1, q}\,;\ s = \overline{1, p})$, and q arbitrarily choosen elements of the l_s order extension of the set $\mathcal{L}$. Being given the operation ε_{rs} between the elements of X_r and X_s that must be implemented by the digital electronic circuit we give all the values to the variables $x \in X_r$, $y \in X_s$ and, through the coding functions adopted $\tilde{f}_r$ respectively $\tilde{f}_s$ we establish the relations through the operation Σ_{rs} between the elements of the extension corresponding to relation (25).

The digital electronic circuit will have the input variables $x_1, x_2, \ldots, x_{l_r}$, $y_1, y_2, \ldots, y_{l_s}$ and the output variables $z_1, z_2, \ldots, z_{l_s}$.

For each output variable we make a Veitch-Karnaugh diagram or any other minimizing method, taking into account the relations obtained through the operation.

For $n < 2^{l_r}$ and $q < 2^{l_s}$ we shall obtain

$$N_{sr} = 2^{l_r + l_s} - nq \tag{28}$$

indifferent states.

4. Concluding Remarks

It is proposed a method for implementation of digital electronic circuits that can perform different operations defined on sets with a finite number of elements. The method is based on the izomorphism in coding.

According to the desired delay, number and type of available logical ports, available signals in the system, etc., one adopts a corresponding coding bijective function. The method is general and efficient.

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POSIBILITATEA SINTEZEI BLOCURILOR DE CALCUL DIGITAL PRIN FOLOSIREA IZOMORFISMULUI ÎN CODARE

(Rezumat)

Se expune o metodă de sinteză a blocurilor de calcul digital ce pot realiza diverse operaţii definite pe mulţimi finite. Funcţie de întîrzierea de propagare impusă prin proiectare şi tipul porţilor logice disponibile rezultă o funcţie de codare bijectivă, după care se implementează blocurile de calcul digital.

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THE STUDY OF THE THREE PHASE A.C. CONTROLLER WITH THREE S.C.R'S FOR A SYMMETRIC INDUCTIVE RECEIVER

BY

M. LUCANU, VALERIU MUNTEANU and AL. HRIŢCU

1. Introduction

In the case of the symmetrical three-phase star-connected receiver, with accessible null, a static switch can be used, made with only three thyristors, whose force-circuit is given in Fig. 1 [1], [2], [4]. By using phase-controlled thyristors. the r.m.s. value of the voltage across the load impedance can be changed, thus resulting an a.c. controller.

In [2] the behaviour of the a.c. controller is studied, in the case of a symmetrical resistive receiver. In what follows we present a method to calculate the currents passing through the load impedance and the voltages across their terminals, if the receiver has an inductive character. This is the most frequent case met in practice.

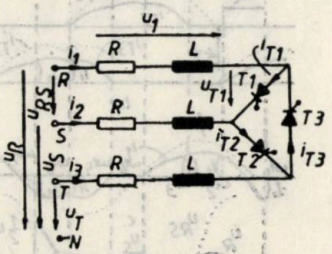

Fig. 1

2. The Waveforms of the Currents and Voltages

In Fig. 2, the waveforms of the voltages of the three-phase supply network, are given, the notations being those used in Fig. 1. Besides, we noted by i_1', i_2', i_3' the currents through the load impedances in the case were the thyristors are controlled with an angle $\alpha \leqslant \varphi$. The angle φ results from the relation :

$$\text{tg } \varphi = \frac{\omega L}{R}, \tag{1}$$

with ω — the angular frequency of the supply alternative voltage.

The control angle α is evalueted from the moment when the voltage across the thyristor (at a purely resistive load) becomes direct, till this is controlled [4].

In Fig. 2 *a* we marked by dark horizontal lines the highest conducting intervals of the three thyristors (which result for $\alpha \leqslant \varphi$). By u_{GK} we noted the impulses which determine the direct commutation of the thyristors.

The waveforms from Fig. 2 are drawn for the case when there are time intervals where two thyristors are simultaneously conducting, i.e. when $\alpha-\varphi \leqslant 2\pi/3$.

In Fig. 2 *b* the waveforms of the currents through the thyristors are represented. We noted by $i'_{T1}, \ldots, i'_{T3}$ the currents through the thyristors when $\alpha \leqslant \varphi$, i.e. when the load impedance are crossed by the currents $i'_1, \ldots, i'_3$.

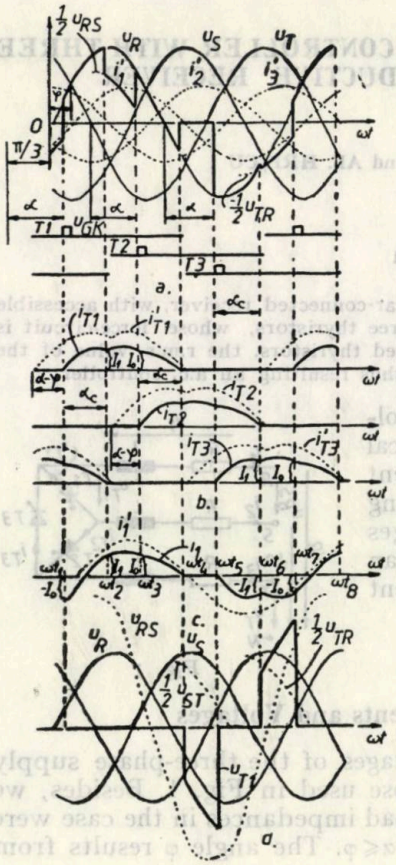

Fig. 2

In the moment t_1, the thyristor $T1$ is comanded to switch-directly, and within the interval $[t_1, t_2]$, the thyristors $T2$ and $T3$ are simultaneously conducting. As a result, the voltage u_1 across the first load impedance (represented in Fig. 2 *a* with a thickened continuous line) will be equal to u_R.

At the moment t_2 the current through $T3$ is zero and, within the interval $[t_2, t_3]$ only $T1$ will be conductive. As a result, the voltage u_1 will be only half the voltage u_{RS}. α_c denotes the angle for which the thyristor $T3$ is still conducting, after the thyristor $T1$ was controlled.

At the moment t_3 the thyristor $T2$ is controlled, within the interval $[t_3, t_4]$, $T1$ and $T2$ are simultaneously conducting and $u_1=u_R$.

Within the interval $[t_4, t_5]$ only the thyristor $T2$ is conductive and hence $u_1=0$. At the moment t_5, the thyristor $T3$ is directly switched, and within the interval $[t_5, t_6]$, $T2$ and $T3$ are conductive and $u_1=u_R$.

Within the interval $[t_6, t_7]$ only the thyristor $T3$ is conductive, resulting $u_1=-u_{TR}/2$.

Taking into account the above considerations and choosing as a origin of time in the point O (Fig. 2 *a*), the voltage across the load impedances will be given by the relations :

$$(2)\qquad u_1(\omega t)=\begin{cases}\sqrt{2}E\sin\omega t, & \text{for } \omega t\in[\omega t_1,\omega t_2]\cup[\omega t_3,\omega t_4]\cup[\omega t_5,\omega t_6],\\ \dfrac{\sqrt{6}E}{2}\sin\left(\omega t+\dfrac{\pi}{6}\right), & \text{for } \omega t\in(\omega t_2,\omega t_3),\\ 0, & \text{for } \omega t\in(\omega t_4,\omega t_5),\\ \dfrac{\sqrt{6}E}{2}\sin\left(\omega t-\dfrac{\pi}{6}\right), & \text{for } \omega t\in(\omega t_6,\omega t_7).\end{cases}$$

By using the relations (2), it results :

$$\bar{U}_1=\frac{1}{2\pi}\int_{-\frac{\pi}{3}+\alpha}^{-\frac{\pi}{3}+\alpha+2\pi} u_1(\omega t)\,\mathrm{d}(\omega t)=0. \tag{3}$$

Since the average value of the voltage across the terminals of one of the load impedances is zero, it follows that the controller can also be used for inductive loads.

In Fig.2 *c*, the waveform of the current i_1 is represented, through the load impedance, which results from the waveforms of the currents through the thyristors and from the relations :

$$i_1=\begin{cases} i_{T1}-i_{T3}, & \text{for } \omega t\in[\omega t_1,\omega t_2],\\ i_{T1}, & \text{for } \omega t\in[\omega t_2,\omega t_4],\\ 0, & \text{for } \omega t\in[\omega t_4,\omega t_5],\\ i_{T3}, & \text{for } \omega t\in[\omega t_5,\omega t_7]. \end{cases} \tag{4}$$

In Fig. 2 *d*, the waveform of the voltage across the thyristor $T1$ is represented. It was drawn on the basis of the following observations :

a) if $T1$ is conductive or $T2$ and $T3$ are simultaneously conductive :

$$u_{T1}=0\ ; \tag{5}$$

b) if only $T2$ is conductive :

$$u_{T1}=u_{RS}+u_{ST}/2\ ; \tag{6}$$

c) if only $T3$ is conductive :

$$u_{T1}=u_{TR}/2+u_{RS}. \tag{7}$$

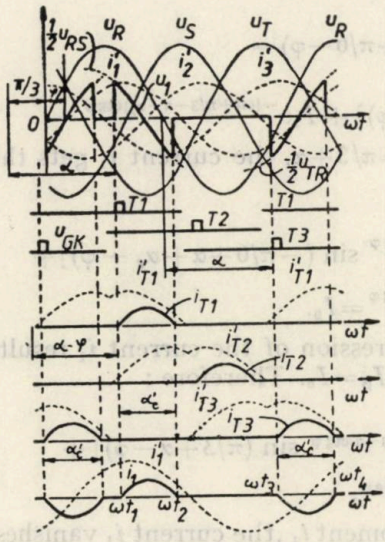

Fig. 3

In Fig. 3, the same waveforms are drawn, but for the case when two thyristors are not simultaneously conductive, i.e. when $\alpha-\varphi\geqslant 2\pi/3$.

The waveforms of the currents result taking into account that, when one thyristor is conductive, two series load impedances are connected at the line voltage.

The waveform of the voltage u_1 is obtained from the relations :

$$u_1(\omega t)=\begin{cases} \frac{1}{2}u_{RS}=\frac{\sqrt{6}E}{2}\sin\left(\omega t+\frac{\pi}{6}\right), \\ \quad \text{for } \omega t\in[\omega t_1,\omega t_2], \text{ when } T1 \text{ is conductive,} \\ -\frac{1}{2}u_{TR}=\frac{\sqrt{6}E}{2}\sin\left(\omega t-\frac{\pi}{6}\right), \\ \quad \text{for } \omega t\in[\omega t_3,\omega t_4], \text{ when } T3 \text{ is conductive,} \\ 0, \text{ on the rest time interval.} \end{cases} \tag{8}$$

3. The Calculation of the Load Currents and Voltages

We shall analyse, at first, the case where there are time intervals when two thyristors are simultaneously conducting. For $\omega t \in [\omega t_1, \omega t_2] \cup [\omega t_3, \omega t_4] \cup [\omega t_5, \omega t_6]$, the expression of the current i_1 is obtained by integrating the differential equation :

$$L\frac{di_1}{dt} + Ri_1 = \sqrt{2}E \sin \omega t, \tag{9}$$

with the initial condition $i_1(\omega t = \beta) = I_\beta$, which gives [3, p. 65] :

$$i_1(\omega t) = \frac{\sqrt{2}E}{R}\cos\varphi[\sin(\omega t - \varphi) - e^{-(\omega t-\beta)\operatorname{ctg}\varphi}\sin(\beta - \varphi)] + I_\beta e^{-(\omega t-\beta)\operatorname{ctg}\varphi}. \tag{10}$$

Within the interval $\omega t \in [\omega t_1, \omega t_2]$, there are the initial conditions $\beta = -\pi/3 + \alpha$, $I_\beta = -I_0$ (Fig. 2 *c*) which, introduced in Eq. (10), give :

$$i_1(\omega t) = \frac{\sqrt{2}E}{R}\cos\varphi[\sin(\omega t - \varphi) - e^{-(\omega t+\pi/3-\alpha)\operatorname{ctg}\varphi}\sin(-\pi/3 + \alpha - \varphi)] - I_0 e^{-(\omega t+\pi/3-\alpha)\operatorname{ctg}\varphi}. \tag{11}$$

For the interval $\omega t \in [\omega t_2, \omega t_3]$, the current i_1 is obtained by integrating the differential equations :

$$L\frac{di_1}{dt} + Ri_1 = \frac{\sqrt{6}E}{2}\sin\left(\omega t + \frac{\pi}{6}\right), \tag{12}$$

with the initial condition $i_1(\omega t = -\pi/3 + \alpha + \alpha_c) = I_1$, which gives :

$$i_1(\omega t) = \frac{\sqrt{6}E}{2R}\cos\varphi\,[\sin(\omega t + \pi/6 - \varphi) - e^{-(\omega t+\pi/3-\alpha-\alpha_c)\operatorname{ctg}\varphi}\sin(-\pi/6 + \alpha + \alpha_c - \varphi)] + I_1 e^{-(\omega t+\pi/3-\alpha-\alpha_c)\operatorname{ctg}\varphi}. \tag{13}$$

At the end of this interval, i.e. for $\omega t = \pi/3 + \alpha$, the current i_1 gets the value I_0. Therefore :

$$\frac{\sqrt{6}E}{2R}\cos\varphi[\sin(\pi/2 + \alpha - \varphi) - e^{-(2\pi/3-\alpha_c)\operatorname{ctg}\varphi}\sin(-\pi/6 + \alpha + \alpha_c - \varphi)] + I_1 e^{-(2\pi/3-\alpha_c)\operatorname{ctg}\varphi} = I_0. \tag{14}$$

For the interval $\omega t \in [\omega t_3, \omega t_4]$, the expression of the current i_1 results by substituting in Eq. (10) $\beta = \pi/3 + \alpha$ and $I_\beta = I_0$. Therefore :

$$i_1(\omega t) = \frac{\sqrt{2}E}{R}\cos\varphi[\sin(\omega t - \varphi) - e^{-(\omega t-\pi/3-\alpha)\operatorname{ctg}\varphi}\sin(\pi/3 + \alpha - \varphi)] + I_0 e^{-(\omega t-\pi/3-\alpha)\operatorname{ctg}\varphi}. \tag{15}$$

At the end of this interval, i.e. at the moment t_4, the current i_1 vanishes. Hence in Eq. (15) $i_1(\omega t = \pi/3 + \alpha + \alpha_c) = 0$. From this condition, after some intermediary calculations, it follows :

$$I_0=\frac{\sqrt{2}E}{R}\cos\varphi[\sin(\pi/3+\alpha-\varphi)-e^{\alpha_c\operatorname{ctg}\varphi}\sin(\pi/3+\alpha+\alpha_c-\varphi)]. \tag{16}$$

Within the interval $\omega t\in[\omega t_4,\omega t_5]$, it results :

$$i_1(\omega t)=0. \tag{17}$$

The expression of the current $i_1(\omega t)$ in the interval $\omega t\in[\omega t_5,\omega t_6]$ is obtained by substituting in Eq. (10) $\beta=\pi+\alpha$ and $I_\beta=0$, whence :

$$i_1(\omega t)=\frac{\sqrt{2}E}{R}\cos\varphi[\sin(\omega t-\varphi)+e^{-(\omega t-\pi-\alpha)\operatorname{ctg}\varphi}\sin(\alpha-\varphi)]. \tag{18}$$

Since at the end of this interval, i.e. in the moment t_6, the current through the load impedance is $-I_1$, in Eq. (18) it follows $i_1(\omega t=\pi+\alpha+\alpha_c)=-I_1$. For this particular case, the expression of the current I_1 has the form :

$$I_1=\frac{\sqrt{2}E}{R}\cos\varphi[\sin(\alpha+\alpha_c-\varphi)-e^{-\alpha_c\operatorname{ctg}\varphi}\sin(\alpha-\varphi)]. \tag{19}$$

Within the interval $\omega t\in[\omega t_6,\omega t_7)]$, the current i_1 will be found by integrating the differential equation

$$L\frac{di_1}{dt}+Ri_1=\frac{\sqrt{6}E}{2}\sin\left(\omega t-\frac{\pi}{6}\right) \tag{20}$$

with the initial condition $i_1(\omega t=\pi+\alpha+\alpha_c)=-I_1$, which gives :

$$i_1(\omega t)=\frac{\sqrt{6}E}{2R}\cos\varphi\,[\sin(\omega t-\pi/6-\varphi)- \tag{21}$$
$$-e^{-(\omega t-\pi-\alpha-\alpha_c)\operatorname{ctg}\varphi}\sin(5\pi/6+\alpha+\alpha_c-\varphi)]-I_1 e^{-(\omega t-\pi-\alpha-\alpha_c)\operatorname{ctg}\varphi}.$$

The angle α_c will be found by introducing (16) and (19) in Eq. (14). After some intermediary calculations, the transcendental equation become :

$$\sin(\alpha+\pi/3+\alpha_c-\varphi)=\frac{e^{-(2\pi/3)\operatorname{ctg}\varphi}+1/2}{e^{-(2\pi/3)\operatorname{ctg}\varphi}/2+1}e^{-\alpha_c\operatorname{ctg}\varphi}\sin(\alpha-\varphi). \tag{22}$$

Since the right member of the relation (22) is always positive, it results that the equations deduced up to now are true as long as is true the inequality

$$\alpha+\alpha_c-\varphi<\frac{2\pi}{3}. \tag{23}$$

Under these circumstances, taking into account the relations (2), the r.m.s. value of the voltage across the load impedance, U_1, can be calculated from the relation :

$$U_1^2=\frac{1}{2\pi}\Bigg[\int_{-\frac{\pi}{3}+\alpha}^{-\frac{\pi}{3}+\alpha+\alpha_e}2E^2\sin^2\omega t\,\mathrm{d}(\omega t)+\int_{-\frac{\pi}{3}+\alpha+\alpha_e}^{\frac{\pi}{3}+\alpha}\frac{3E^2}{2}\sin^2(\omega t+\pi/6)\,\mathrm{d}(\omega t)+$$

$$+\int_{\frac{\pi}{3}+\alpha}^{\frac{\pi}{3}+\alpha+\alpha_e}2E^2\sin^2\omega t\,\mathrm{d}(\omega t)+\int_{\pi+\alpha}^{\pi+\alpha+\alpha_e}2E^2\sin^2\omega t\,\mathrm{d}(\omega t)+ \tag{24}$$

$$+\int_{\pi+\alpha+\alpha_e}^{2\pi-\pi/3+\alpha}\frac{3E^2}{2}\sin^2\left(\omega t-\frac{\pi}{6}\right)\mathrm{d}(\omega t)\Bigg],$$

or :

$$U_1=\frac{E}{2}\sqrt{\frac{1}{\pi}\left[3\alpha_e+2\pi+\frac{3}{2}\sin 2\alpha-\frac{3}{2}\sin 2\left(\alpha+\alpha_e+\frac{\pi}{3}\right)\right]}. \tag{25}$$

If $\alpha-\varphi\geqslant 2\pi/3$, there will no longer exist time intervals where two thyristors will simultaneously be conductive. The waveforms given in Fig. 3 corresponds to this a situation.

Within the interval $\omega t\in[\omega t_1,\ \omega t_2]$, the current $i_1(\omega t)$ results by integrating the differential Eq. (12), with the initial condition $i_1(\omega t=\pi/3+\alpha)=0$, which gives the expression :

$$i_1(\omega t)=\frac{\sqrt{6}E}{2R}\cos\varphi[\sin(\omega t+\pi/6-\varphi)-\mathrm{e}^{-(\omega t+\pi/3-\alpha)\operatorname{ctg}\varphi}\sin(\pi/6+\alpha-\varphi)]. \tag{26}$$

In the interval $\omega t\in[\omega t_2,\ \omega t_3]$, the current $i_1(\omega t)$ is zero.

The current $i_1(\omega t)$ for the interval $\omega t\in[\omega t_3,\ \omega t_4]$ will be obtained by integrating the differential Eq. (20), with the initial condition $i_1(\omega t=\pi+\alpha)=0$ resulting :

$$i_1(\omega t)=\frac{\sqrt{6}E}{2R}\cos\varphi\,[\sin(\omega t-\pi/6-\varphi)-\mathrm{e}^{-(\omega t-\pi-\alpha)\operatorname{ctg}\varphi}\sin(5\pi/6+\alpha-\varphi]. \tag{27}$$

The condition angle α_e' of a thyristor can be determined by imposing in Eq. (26) the condition $i_1(\omega t=-\pi/3+\alpha+\alpha_e')=0$, whence the following transcendental equation will be obtained :

$$\sin(-\pi/6+\alpha+\alpha_e'-\varphi)=\mathrm{e}^{-\alpha_e'\operatorname{ctg}\varphi}\sin(-\pi/6+\alpha-\varphi). \tag{28}$$

For a concrete determination of the angle α_e', one can use the nomograms from [5, p. 8], with the specification that the abscissae and the ordinates must be properly prelucrated.

For the studied case, taking into account the relations (8), the r.m.s. value of the voltage across a load impedance is

$$(29)\qquad U_1^2=\frac{1}{2\pi}\left[\int_{-\frac{\pi}{3}+\alpha}^{-\frac{\pi}{3}+\alpha+\alpha_c'}\frac{3E^2}{2}\sin^2(\omega t+\pi/6)\mathrm{d}(\omega t)+\int_{\pi+\alpha}^{\pi+\alpha+\alpha_c'}\frac{3E^2}{2}\sin^2(\omega t-\pi/6)\mathrm{d}(\omega t)\right],$$

or :

$$(30)\qquad U_1=\frac{E}{2}\sqrt{\frac{3}{\pi}[\alpha_c'+\sin\alpha_c'\cos(-\pi/3+2\alpha+\alpha_c')]}.$$

4. Determination of the Currents Passing through the Thyristors

We shall initially consider the case when $\alpha-\varphi<2\pi/3$, i.e. there are time intervals when two thyristors are simultaneously conducting.

Only the expresions for the current through the thyristor $T1$ will be deduced, the currents through the thyristors $T2$ and $T3$ being identical, only dephased with $2\pi/3$ and $4\pi/3$, respectively.

Refering to Figs. 1 and 2 and choosing the origin of time also in the point O, for the time interval $\omega t\in[\omega t_1, \omega t_2]$, the current through the thyristor $T1$ will be $i_{T1}=-i_2$. Therefore, it will be obtained by integrating the differential Eq. :

$$(31)\qquad L\frac{\mathrm{d}i_{T1}}{\mathrm{d}t}+Ri_{T1}=-\sqrt{2}E\sin(\omega t-2\pi/3)=\sqrt{2}E\sin(\omega t+\pi/3),$$

with the initial condition $i_{T1}(\omega t=-\pi/3+\alpha)=0$, resulting :

$$(32)\qquad i_{T1}(\omega t)=\frac{\sqrt{2}E}{R}\cos\varphi[\sin(\omega t+\pi/3-\varphi)-\mathrm{e}^{-(\omega t+\pi/3-\alpha)\operatorname{ctg}\varphi}\sin(\alpha-\varphi)].$$

For the time interval $\omega t\in[\omega t_2, \omega t_3]$, the current i_{T1} is :

$$(33)\qquad i_{T1}(\omega t)=i_1(\omega t)=\frac{\sqrt{6}E}{2R}\cos\varphi[\sin(\omega t+\pi/6-\varphi)-$$
$$-\mathrm{e}^{-(\omega t+\pi/3-\alpha-\alpha_c)\operatorname{ctg}\varphi}\sin(-\pi/6+\alpha+\alpha_c-\varphi)]+I_1\mathrm{e}^{-(\omega t+\pi/3-\alpha-\alpha_c)\operatorname{ctg}\varphi}.$$

Similarly, for the interval $\omega t\in[\omega t_3, \omega t_4]$ we can write the relation :

$$(34)\qquad i_{T1}(\omega t)=i_1(\omega t)=\frac{\sqrt{2}E}{R}\cos\varphi[\sin(\omega t-\varphi)-$$
$$-\mathrm{e}^{-(\omega t-\pi/3-\beta)\operatorname{ctg}\alpha}(\sin\pi/3+\alpha-\varphi)]+I_0\mathrm{e}^{-(\omega t-\pi/3-\alpha)\operatorname{ctg}\varphi}.$$

For the rest of the time interval, the current through the thyristor $T1$ is zero.

If $\alpha-\varphi\geqslant 2\pi/3$, at a given moment we can only have one conducting thyristor, as can be seen from the waveforms from Fig. 3. In this case, for $\omega t\in[\omega t_1, \omega t_2]$, the current through the thyristor $T1$ is given by the relation :

$$(35)\qquad i_{T1}(\omega t)=i_1(\omega t)=\frac{\sqrt{6}E}{2R}\cos\varphi[\sin(\omega t+\pi/6-\varphi)-$$
$$-\mathrm{e}^{-(\omega t+\pi/3-\alpha)\operatorname{ctg}\varphi}\sin(-\pi/6+\alpha-\varphi)],$$

this being zero all along the rest of the interval.

5. Conclusions

The analysis of a force-circuit of a three phase A.C. controller, performed with three thyristors, having a symmetrical inductive receiver, is made.

In a particular case, given the load parameters R, L i.e. the phase difference φ, the determination of the load currents and voltages is accomplished in the following way:

a) If $\alpha-\varphi<2\pi/3$, by using the relation (22) the angle α_c will be determined. With the relations (16) and (19) we can determine I_0 and I_1. By replacing their values in the relations (11), (13), (15), (18) and (21) the expression of the load current will be obtained within the following intervals $[\omega t_1, \omega t_2]$, $[\omega t_2, \omega t_3]$, $[\omega t_3, \omega t_4]$, $[\omega t_5, \omega t_6]$, $[\omega t_6, \omega t_7]$. The r.m.s. value of the load voltage can be calculated with the relation (25).

b) If $\alpha-\varphi\geqslant 2\pi/3$, the relation (28) will be used for deducing the angle α_c'. The load current $i_1(\omega t)$ will be determined from the relations (26) and (27) for the intervals $[\omega t_1, \omega t_2]$ and $[\omega t_3, \omega t_4)]$ respectively, and the value of the load voltage from the relation (30).

It follows from Fig. 2 *d* that for this controller, in the case of inductive loads, there are also moments $(t=t_6)$ when on the thyristors are applied direct anodic voltages with very high increasing rates, the thyristors demanding a proper selection.

For particular cases, from the relations deduced for $\varphi=0$ the results from [2] will be obtained.

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STUDIUL VARIATORULUI TRIFAZAT DE TENSIUNE ALTERNATIVĂ CU TREI TIRISTOARE, AVÎND UN RECEPTOR INDUCTIV SIMETRIC

(Rezumat)

Se efectuează analiza circuitului de forţă al unui variator trifazat de tensiune alternativă realizat cu trei tiristoare, avînd un receptor inductiv simetric. Se deduc formele de undă şi relaţiile de calcul ale curenţilor şi tensiunilor de sarcină.

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D.K. 611.313.3

ÜBER EIN PRINZIP, SCHWINGUNGEN VON SCHRITTMOTOREN ZU DÄMPFEN

VON

CRISTEA PAL

1. Einführung

Bei einem elektrischen Betriebssystem erfüllt ein Schrittmotor gleichzeitig zwei Funktionen: er wandelt elektrische Energie in mechanische Energie um und überträgt eine ziffernmassige elektrische. Information in eine angulare oder lineare Position. Die zweite Funktion bedingt es, daß die Leistungen des Antriebsystems und im besondern die dynamischen Leistungen eng an den Typus des Zuführungs- und Steuerschemas gebunden sind [1], [2], besonders bei Systemen, die in offenem Kreis arbeiten.

In der Arbeit werden Beiträge über die Dämpfung von Schwingungen des beweglichen Motorenteiles im Falle quasistationärer Betriebsweisen erbracht.

2. Dämpfung von Schrittmotorenschwingungen bei veränderlicher Reluktanz durch Verzögerung der Steureimpulse

Die vorgeschlagene Dämpfungsmethode kann an Schrittmotoren mit veränderlicher Reluktanz angewendet werden, bei welchen mehrere identisch in Bezug auf eine Beziehungsachse montierte Statorenpakete und ein einziger mit kollinearen Polteilen versehener Rotor vorhanden sind. Diese Anordnung ist gleicharting mit einer Gruppe identischer Schrittmotore, die parallel geschaltet sind [3].

Es wird der Fall eines Motors mit drei Statorpaketen erläutert, bei welchem der Beitrag jedes Paketes zum Gesamtdrehmoment M des Motors gleich ist mit $M=M_1+M_2+M_3$, wobei $M_1=M_2=M_3$ die Teilmomente vorstellen.

In Abb. 1 sind die Kurven der statischen Momente in Abhängigkeit vom Verschiebungswinkel θ dargestellt u. zw.: Kurve *1* stellt die statischen Teilmomente $M_{10}=M_{20}=M_{30}$ vor, die einer Speisungsfolge entsprechen, für welche das stabile Gleichgewicht des Rotors im Punckt *O* vorliegt, wenn auf die Motorachse ein resistentes Drehmoment $M_r=0$ einwirkt; Kurve *2* stellt das Gesamtdrehmoment M_0 des Motors unter denselben Bedingungen wie bei Kurve *1* vor; Kurve *3* stellt Gesamtdrehmoment M_1 des Motors vor, welches einer neuen Speisungsfolge entspricht, für welche der Motor

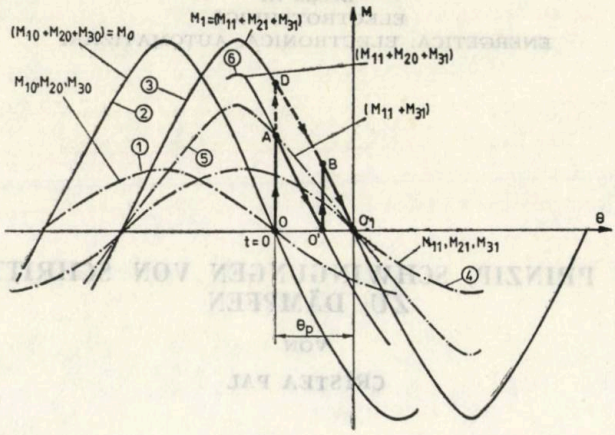

Abb. 1

einen Schritt vom Werte θ_p vollführt hat ; das stabile Gleichgewicht liegt also in O_1 und $M_1=M_{11}+M_{21}+M_{31}$; Kurve *4* stellt die statischen Teilmomente $M_{11}=M_{12}=M_{31}$ vor, die einer neuen Speisungsfolge entsprechen ; Kurve *5* stellt das statische Teilmoment $M_{11}+M_{31}$ für den Fall vor, in welchem der Motor nur die nach der neuen Folge gespeisten Statoren *1* und *3* aufweist ; Kurve *6* stellt das resultierende statische Moment $M_{11}+M_{20}+M_{31}$ für den Fall vor, in welchem der Motor die nach der neuen Folge gespeisten Statoren *1* und *3* und den Stator *2* nach der alten Folge aufweist ; in diesem Fall befindet sich das stabile Betriebsgleichgewicht in O' für ein resistentes Moment an der Welle $M_r=0$.

Wenn die Anregung des erläuterten Schrittmotors nach einer normalen Folge stattfindet und man einem Vorschub von einem Schritt aus des Stellung O nach O_1 wünscht, werden im Zeitpunkte $t=0$ die Motorstatoren so angeregt, daß der Betrieb des Motors von der Charakteristik *2* auf die Charakteristik *3* übergeht. Unter idealen Bedingungen bei Vernachlässigung der elektromagnetischen Zeitkonstanten, induzierter EMK und nichtlinearer magnetischer Phänomene geschieht der Übergang entlang der Geraden OAD. Der Rotor des Motors beschleunigt gegen Punct O_1, speichart also kinetische Energie, die in O_1 einen Höchstwert erreicht. Um den Punkt O_1 werden Schwingungen bis zur Dissipation der aufgespeicherten Energie stattfinden. Die vom Rotor angehäufte Energie ist der Fläche ODO_1 proportionell und ist groß.

Die Dämpfung der Rotorschwingungen kann durch die Verwendung einer Steuerschaltung geschehen, die eine Zeitverzögerung τ_i bei der Übermittlung des Schaltsignales der Speisung der Phasen eines der Statoren des Schrittmotors bedingen.

Wir betrachten als Ausgangspunkt den Punkt des stabilen Gleichgewichtes O und es wird gewünscht, daß der Motor einen Schritt von O nach O_1 ausführe. Dazu regt man die Statoren *1* und *3* gemäß der Folge an, die dem stabilen Gleichgewicht O_1, der Charakteristik *5* entspricht, und hält

für den Stator *2* die dem Gleichgewichtspunkt *O* entsprechende Speisung, bzw. die Charakteristik des Moments *1*. Die Charakteristik des resultierenden statischen Momentes für diesen Fall wird von Kurve *6* gegeben. Der Betriebspunkt geht von *O* nach *A* auf Kurve *6* über und der Motor beschleunigt nach dem stabilen Gleichgewicht *O'*. Bei Berührung dieses Punktes, bzw. bei $t=\tau_1$ oder gegebenenfalls vor oder nach seiner Berührung, schaltet auch Stator *2* gemäß einer Speisungsfolge, die dem stabilen Gleichgewicht in O_1 entspräche. Der Betrieb des Motors geht von *O'* nach *B* auf der Charakteristik *3* und schließlich nach O_1 über. Man beobachtet, daß durch diese Steureung die Fläche *ADBO'*, der Beschleunigung aus der Fläche ODO_1 eliminiert wurde ; der Motor in die Stellung des stabilen Gleichgewichtes mit einer geringern Geschwindigkeit gelangen und die Amplitude der Schwingungen wird abnehmen.

3. Versuchsarbeiten

Die Versuchsarbeiten wurden an einem linearen Schrittmotor von eigenen Bau durchgeführt, welcher den gewünschten Bedingungen entsprich.

In Abb. 2 sind die Charakteristiken Schub—Zeit, $x=f(t)$, für eine Schrittfrequenz von 2,27 Hz dargestellt.

Die Charakteristiken *1* und *2* sind für den ungebremsten Motor $\tau_i=0$, für Leerlauf und mit zusätzlicher Beharrungsmasse von 2 Kg aufgestellt.

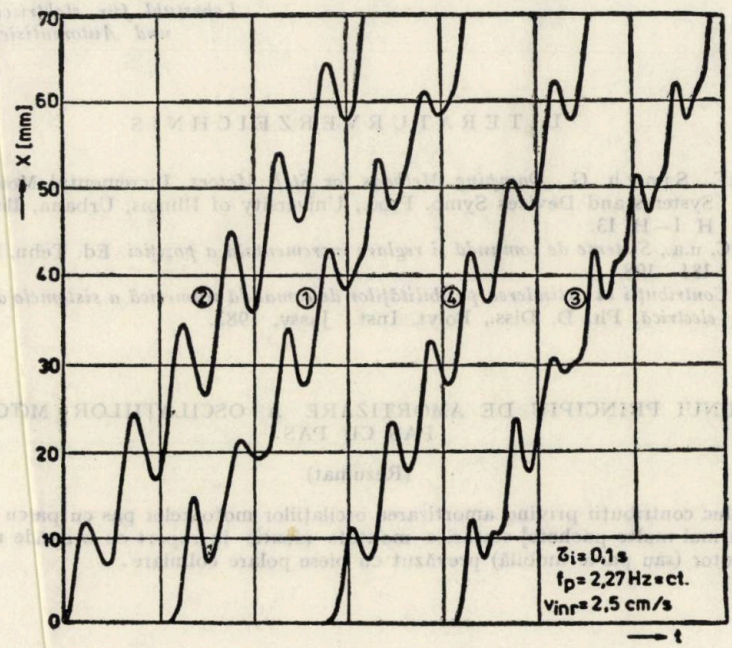

Fig. 2

Bei der Charakteristik *1* beträgt der Mittelwert der Überregelung 29% und für die Charakteristik *2*, 42%. Man stellt ein starkes Ansteigen der Schwingungsamplitude mit der Vergösserung der Beharrungmasse fest.

Die Charakteristiken *3* und *4* sind hoch, beim gebremstem Motor, τ_i=0,1 s, im Leerlauf und mit einer zusätzlichen Beharrungsmasse von 2 Kg. Der Mittelwert der Überregelung für die Charakteristik *4* beträgt 23%, 26% jedoch für die Charakteristik *3*. Die Verzögerung τ_i wurde durch wiederholte Prüfungen festgestellt.

4. Schlußfolgerungen

1. Die vorgeschlagene Dämpfungsmethode von Schwingungen bietet den Vorteil, daß sie bei Motoren anwendbar ist, die mittels klassischer Schaltungen in offenem Kreis gesteuert werden ; die einzige Abänderung besteht in der Einführung von Verzögerungsblöcken der Schaltsignale, welche die Schrittfrequenz bedingen.

2. Gebremst kann bei jeden vom Motor ausgeführten Schritt bzw. nur beim letzten Schritt werden.

3. Die Verzögerung τ_i ist eine Funktion der Natur und Art der Veränderung des resistenten Momentes. Die Methode hat demgemäß eine veränderliche Wirksamkeit je nach den Betriebsbedingungen des Motors und bedingt eine Anpassung an diese Bedingungen bzw. ein Regelung der Zeitdauer τ_i.

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ASUPRA UNUI PRINCIPIU DE AMORTIZARE A OSCILAȚIILOR MOTOARELOR PAS CU PAS

(Rezumat)

Se aduc contribuții privind amortizarea oscilațiilor motoarelor pas cu pas cu reluctanță variabilă cu mai multe pachete statorice montate identic în raport cu o axă de referință și un singur rotor (sau parte mobilă) prevăzut cu piese polare coliniare.

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ECUAŢIILE MAŞINII CU REPULSIE CU COMUTAŢIE STATICĂ COMANDATĂ PE O ALTERNANŢĂ

DE

SILVIU HARGHEL şi EMIL TEODORU

1. Introducere

În [1]...[3] s-au prezentat diverse variante de realizare a maşinii cu repulsie cu comutaţie statică. În toate situaţiile s-a luat în discuţie şi posibilitatea de funcţionare cu comandă pe o singură alternanţă. Această modalitate de lucru se poate utiliza pentru obţinerea unui reglaj suplimentar de viteză sau, în mod special, la schemele maşinilor de mică putere, unde o diminuare moderată a performanţelor se poate compensa printr-un preţ de cost mai redus şi prin dimensiuni mai mici.

În lucrarea de faţă se deduc ecuaţiile de funcţionare ale maşinii susamintite urmîndu-se aceeaşi metodologie de calcul ca în [4].

2. Expresiile curenţilor maşinii

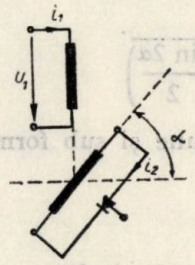

Fig. 1

În fig. 1 se reprezintă schema de principiu a maşinii cu repulsie, cu comandă pe o alternanţă. Elementele de comandă de forţă pot fi, aşa cum s-a admis, tiristoare sau, pentru maşini de mică putere, tranzistoare.

În cele ce urmează s-a considerat elementul redresor comandat ca fiind ideal ($R_d=0$, $R_i=\infty$) şi funcţionînd cu deschidere totală. Se face şi o a doua aproximaţie considerînd curentul secundar i_2 în perioada cînd redresorul conduce, de forma corespunzătoare funcţionării maşinii în sistem clasic.

Sistemul de ecuaţii diferenţiale ce descriu funcţionarea maşinii, la care trebuie ataşată caracteristica redresorului, este de forma:

$$(1)\qquad \begin{cases} u_1 = R_1 i_1 + L_2 \dfrac{di_1}{dt} + M \dfrac{di_2}{dt} \sin\alpha, \\ \Delta u_T = R_2 i_2 + L_2 \dfrac{di_2}{dt} + M \dfrac{di_1}{dt} \sin\alpha - M i_1 \cos\alpha \dfrac{d\alpha}{dt}, \end{cases}$$

care în ipoteza $\Delta u_T = 0$ (căderile de tensiune pe elementele redresoare comandate nule) şi cu notaţia $d\alpha/dt = \Omega$ devin :

$$(2)\quad \begin{cases} u_1 = R_1 i_1 + L_1 \dfrac{di_1}{dt} + M \dfrac{di_2}{dt} \sin\alpha, \\ 0 = R_2 i_2 + L_2 \dfrac{di_2}{dt} + M \dfrac{di_1}{dt} \sin\alpha - M\Omega i_1 \cos\alpha. \end{cases}$$

În baza ipotezei acceptate se consideră că în perioada de conducţie curenţii sînt daţi de relaţiile :

$$(3)\quad i_1 = I_{1m} \sin(\omega t - \varphi_1), \quad i_2 = I_{2m} \sin(\omega t - \varphi_2),$$

unde

$$(4)\quad I_{1m} = \frac{\sqrt{2} U_1 Z_2}{\sqrt{an^2 + bn + c}}, \quad \cos\varphi_1 = \frac{d + gn}{an^2 + bn + c},$$

notaţiile fiind cele utilizate în [4].

Ţinînd cont de expresia curentului secundar al maşinii cu repulsie stabilită în [4] :

$$(5)\quad \underline{I}_2 = \frac{-\underline{U}_1 \underline{Z}_0 \left(\cos\alpha + j\frac{n}{f}\sin\alpha\right)}{\underline{Z}_1\underline{Z}_2 - \underline{Z}_0^2\left(\cos^2\alpha + j\frac{n}{f}\cdot\frac{\sin 2\alpha}{2}\right)},$$

înlocuind impedanţele cu expresiile: $\underline{Z}_k = R_k + jX_k, (k = 1, 2)$, $\underline{Z}_0 = jX_h$, rezultă:

$$(6)\quad \underline{I}_2 = \frac{j\underline{U}_1 X_h \left(\cos\alpha + j\frac{n}{f}\sin\alpha\right)}{(R_1 + jX_1)(R_2 + jX_2) + X_h^2\left(\cos^2\alpha + j\frac{n}{f}\cdot\frac{\sin 2\alpha}{2}\right)}.$$

Prin efectuarea calculelor expresia (6) se poate pune şi sub forma :

$$(7)\quad \underline{I}_2 = U_1 X_h \left(j\cos\alpha - \frac{n}{f}\sin\alpha\right) \times \\ \times \frac{R_1R_2 - X_1X_2 + X_h^2\cos^2\alpha - j\left(X_1R_2 + R_1X_2 + X_h^2\frac{n}{f}\cdot\frac{\sin 2\alpha}{2}\right)}{(R_1R_2 - X_1X_2 + X_h^2\cos^2\alpha)^2 + \left(X_1R_2 + R_1X_2 + X_h^2\frac{n}{f}\cdot\frac{\sin 2\alpha}{2}\right)^2},$$

de unde se deduc relaţiile :

$$(8)\quad I_{2m} = \frac{\sqrt{2} U_1 X_h \sqrt{\cos^2\alpha + \frac{n^2}{f^2}\sin^2\alpha}}{\sqrt{an^2 + bn + c}}$$

și

$$I_2 \sin \varphi_2 = \left[(R_1R_2 - X_1X_2 + X_k^2 \cos \alpha) \cos \alpha + \frac{n}{f} \left(X_1R_2 + R_1X_2 + \right.\right.$$
$$\left.\left. + \frac{n}{f} X_k \frac{\sin 2\alpha}{2} \right) \sin \alpha \right] X_k U_1. \tag{9}$$

Pentru unghiul φ_2 rezultă expresia :

$$\varphi_2 = \operatorname{arc\,tg} \left[(R_1R_2 - X_1X_2 + X_k^2 \cos \alpha) \cos \alpha + \frac{n}{f} \sin \alpha \left(X_1R_2 + R_1X_2 + \right.\right.$$
$$\left.\left. + X_n^2 \frac{n}{f} \cdot \frac{\sin 2\alpha}{2} \right) \right] \Big/ \left[\left(X_1R_2 + R_1X_2 + X_k^2 \frac{n}{f} \cdot \frac{\sin 2\alpha}{2} \right) \cos \alpha - \right.$$
$$\left. - \frac{n}{f} \sin \alpha (R_1R_2 - X_1X_2 + X_k^2 \cos \alpha) \right]. \tag{10}$$

Curentul i_2 va avea deci într-o alternanță expresia $i_2 = I_{2m} \sin (\omega t - \varphi_2)$, cu I_{2m} și φ_2 dați mai sus, iar pentru celalaltă alternanță $i_2 = 0$. Vom considera acest curent periodic, avînd perioada $T = 2\pi/\omega$, încît se poate dezvolta într-o serie Fourier. Așadar

$$i_2 = \begin{cases} I_{2m} \sin (\omega t - \varphi_2), & \text{pentru } 0 < \omega t < \pi, \\ 0, & \text{pentru } \pi < \omega t < 2\pi. \end{cases} \tag{11}$$

Seria Fourier corespunzătoare este de forma :

$$i_2 = \frac{A_0}{2} + A_1 \cos \omega t + A_2 \cos 2\omega t + \ldots + B_1 \sin \omega t + B_2 \sin 2\omega t + \ldots \tag{12}$$

Efectuînd schimbarea de variabilă $\omega t = x$ se calculează coeficienții seriei A_0, A_1 și B_1 cu ajutorul relațiilor cunoscute. Astfel

$$A_0 = \frac{2}{T} \int_0^T f(t)\, dt = \frac{2}{T\omega} \int_{\varphi_2}^{\varphi_2 + \pi} I_{2m} \sin (x - \varphi_2)\, dx. \tag{13}$$

Dacă se ține cont că $\omega T = 2\pi$ după efectuarea integralei se obține

$$A_0 = \frac{2I_{2m}}{\pi}. \tag{14}$$

Pentru obținerea coeficientului A_1 se utilizează relația :

$$A_1 = \frac{2}{T} \int_0^T f(t) \cos \omega t\, dt = \frac{2}{T\omega} \int_{\varphi_2}^{\varphi_2 + \pi} I_{2m} \sin (x - \varphi_2) \cos x\, dx \tag{15}$$

de unde rezultă :

$$A_1 = -\frac{I_{2m}}{2}\sin\varphi_2. \tag{16}$$

În final pentru coeficientul B_1 se folosește expresia :

$$B_1 = \frac{2}{T}\int_0^T f(t)\sin\omega t\,dt = \frac{2}{T\omega}\int_{\varphi_2}^{\varphi_2+\pi} I_{2m}\sin(x-\varphi_2)\sin x\,dx, \tag{17}$$

obținînd

$$B_1 = \frac{I_{2m}}{2}\cos\varphi_2. \tag{18}$$

Reținînd din dezvoltarea în serie numai componenta continuă și termenul fundamental și ținînd cont de expresiile (14), (16) și (18) rezultă pentru curentul secundar i_2 forma aproximativă :

$$i_2 = \frac{I_{2m}}{\pi} - \frac{I_{2m}}{2}\sin\varphi_2\cos\omega t + \frac{I_{2m}}{2}\cos\varphi_2\sin\omega t, \tag{19}$$

unde, pentru simplificarea calculelor ce urmează, se introduc următoarele notații :

$$I_0 = \frac{I_{2m}}{\pi},\quad I_c = -\frac{I_{2m}}{2}\sin\varphi_2,\quad I_s = \frac{I_{2m}}{2}\cos\varphi_2. \tag{20}$$

Curentul i_2 devine, în aceste condiții, de forma :

$$i_2 = I_0 + I_c\cos\omega t + I_s\sin\omega t. \tag{21}$$

Reluăm prima ecuație din sistemul (1) în care se consideră rezistența înfășurării primare neglijabilă ($R_1 = 0$) :

$$u_1 = L_1\frac{di_1}{dt} + M\frac{di_2}{dt}\sin\alpha. \tag{22}$$

Înlocuind $u_1 = U_{2m}\sin\omega t$ și expresia (21) a lui i_2 în ecuația (22) se obține :

$$U_{1m}\sin\omega t = L_1\frac{di_1}{dt} + M\sin\alpha\,(-\omega I_c\sin\omega t + \omega I_s\cos\omega t), \tag{23}$$

de unde rezultă

$$\frac{di_1}{dt} = \frac{U_{1m}}{L_1}\sin\omega t + \omega\frac{M}{L_1}I_c\sin\alpha\sin\omega t - \omega\frac{M}{L_1}I_s\sin\alpha\cos\omega t. \tag{24}$$

Integrînd relația (24) se obține expresia curentului i_1 :

$$i_1 = -\frac{U_{1m}}{\omega L_1}\cos\omega t - \frac{M}{L_1}I_c\sin\alpha\cos\omega t - \frac{M}{L_1}I_s\sin\alpha\sin\omega t. \tag{25}$$

3. Expresia cuplului electromagnetic

Pentru determinarea expresiei cuplului electromagnetic se pornește de la forma generală a acestuia :

(26) $$c=\left(\frac{\partial W_m}{\partial \alpha}\right)_{i=ct.},$$

unde W_m reprezintă energia înmagazinată în cîmpul magnetic, al mașinii :

(27) $$W_m=\frac{1}{2}L_1 i_1^2+\frac{1}{2}L_2 i_2^2+m i_1 i_2$$

iar α — unghiul de rotație al rotorului. Considerînd L_1 și L_2 constante rezultă că numai cel de al treilea termen al expresiei contribuie la producerea unui lucru mecanic. Ținînd seama în plus (v. fig. 1) că

(28) $$m=M \sin \alpha$$

rezultă :

(29) $$c=i_1 i_2 \frac{\mathrm{d}m}{\mathrm{d}\alpha}=-i_1 i_2 M \cos \alpha.$$

Introducînd în (29) expresiile curenților i_1 și i_2 date de relațiile (21) și respectiv (25) se obține

(30) $$c=\left(\frac{U_{1m}}{\omega L_1}\cos \omega t+\frac{M}{L_1}I_c \sin \alpha \cos \omega t+\frac{M}{L_1}I_s \sin \alpha \sin \omega t\right)\times$$
$$\times (I_0+I_c \cos \omega t+I_s \sin \omega t)\, M \cos \alpha.$$

Efectuînd calculele și introducînd notația $X_1=\omega L_1$ se obține în final :

$$c=\frac{U_{1m}I_c M \cos \alpha}{2X_1}+\frac{M^2 I_c^2 \sin \alpha \cos \alpha}{2L_1}+\frac{M^2 I_s^2 \sin \alpha \cos \alpha}{2L_1}+$$
$$+\left(\frac{U_{1m}I_0 M \cos \alpha}{X_1}+\frac{M^2}{L_1}I_c I_0 \sin \alpha \cos \alpha\right)\cos \omega t+$$
(32) $$+\frac{M^2}{L_1}I_s I_0 \sin \alpha \cos \alpha \sin \omega t+\left(\frac{U_{1m}I_s M \cos \alpha}{2X_1}+\frac{M^2 I_c^2 \sin \alpha \cos \alpha}{2L_1}-\right.$$
$$\left.-\frac{M^2 I_s^2 \sin \alpha \cos \alpha}{2L_1}\right)\cos 2\omega t+\left(\frac{U_{1m}I_s M \cos \alpha}{2X_1}+\right.$$
$$\left.+\frac{M^2}{L_1}I_c I_s \sin \alpha \cos \alpha\right)\sin 2\omega t.$$

Se introduce notația :

$$C_0=\frac{U_{1m}I_c M \cos \alpha}{2X_1}+\frac{M^2 I_c^2 \sin \alpha \cos \alpha}{2L_1}+\frac{M^2 I_s^2 \sin \alpha \cos \alpha}{2L_1}=$$
$$=\frac{U_{1m}I_c M \cos \alpha}{2X_1}+\frac{M^2 \sin 2\alpha}{4L_1}(I_c^2+I_s^2).$$

Dar

$$I_c^2+I_s^2=\frac{I_{2m}^2}{2}\sin^2\varphi_2+\frac{I_{2m}^2}{2}\cos^2\varphi_2=\frac{I_{2m}^2}{2} \tag{33}$$

încît

$$C_0=\frac{M^2I_{2m}^2\sin 2\alpha}{8L_1}-\frac{MU_{1m}I_{2m}\sin\varphi_2\sin 2\alpha}{4X_1}. \tag{34}$$

De asemenea se mai introduc și notațiile :

$$C_{c1}=\frac{U_{1m}I_0M\cos\alpha}{X_1}+\frac{M^2}{L_1}I_cI_0\sin\alpha\cos\alpha= \tag{35}$$

$$=\frac{U_{1m}I_{2m}M\cos\alpha}{\pi X_1}-\frac{M^2I_{2m}^2\sin\varphi_2\sin 2\alpha}{4\pi L_1},$$

$$C_{s1}=\frac{M^2}{L_1}I_sI_0\sin\alpha\cos\alpha=\frac{M^2I_{2m}^2\cos\varphi_2\sin 2\alpha}{4\pi L_1}, \tag{36}$$

$$C_{c2}=\frac{U_{1m}I_cM\cos\alpha}{2X_1}+\frac{M^2I_c^2\sin\alpha\cos\alpha}{2L_1}-\frac{M^2I_s^2\sin\alpha\cos\alpha}{2L_1}= \tag{37}$$

$$=\frac{M^2I_{2m}^2\sin 2\alpha}{16L_1}(1-2\cos^2\varphi_2)-\frac{U_{1m}I_{2m}M\sin\varphi_2\cos\alpha}{4X_1},$$

$$C_{s2}=\frac{U_{1m}I_sM\cos\alpha}{2X_1}+\frac{M^2}{L_1}I_cI_s\sin\alpha\cos\alpha= \tag{38}$$

$$=\frac{U_{1m}I_{2m}M\cos\varphi_2\cos\alpha}{4X_1}-\frac{M^2}{16L_1}I_{2m}^2\sin 2\varphi_2\sin 2\alpha.$$

Ținînd cont de (34) ... (38) expresia (32) a cuplului electromagnetic capătă forma :

$$c=C_0+C_{c1}\cos\omega t+C_{s1}\sin\omega t+C_{c2}\cos 2\omega t+C_{s2}\sin 2\omega t. \tag{39}$$

4. Expresia cuplului mediu. Concluzii

După cum se observă din relațiile stabilite mașina prezintă un cuplu electromagnetic cu o valoare medie nenulă. Mărimea cuplului electromagnetic depinde de tensiunea de alimentare, de unghiul de decalare a periilor α (de poziție a traductorului rotoric) și de viteza de rotație (i_2 fiind funcție de aceasta). Cuplul mașinii are o componentă de pulsație nulă și componente de pulsații ω și 2ω.

Pentru a obține expresia cuplului mediu se integrează expresia (39) pe un interval de timp egal cu o perioadă :

$$C_{med}=\frac{1}{T}\int_0^T c\,dt=\frac{M^2I_{2m}^2\sin 2\alpha}{8L_1}-\frac{U_{1m}I_{2m}M\sin\varphi_2\cos\alpha}{4X_1}. \tag{40}$$

Dacă în continuare se ține cont de relațiile (7) și (9) rezultă :

$$C_{med} = -\frac{U_1^2 X_h^2 M^2 \left(\cos^2\alpha + \frac{n^2}{f^2}\sin^2\alpha\right)\sin 2\alpha}{8(an^2+bn+c)L_1} -$$

$$-\frac{U_1^2 X_h \left[(R_1R_2 - X_1X_2 + X_h^2\cos\alpha)\cos\alpha}{2X_1} + \right. \tag{41}$$

$$\left. + \frac{\frac{n}{f}\left(X_1R_2 + R_1X_2 + \frac{n}{f}X_h^2\frac{\sin 2\alpha}{2}\right)\sin\alpha\right]\cos\alpha}{2X_1}.$$

Se remarcă și de data aceasta dependența cuplului mediu al mașinii de tensiunea de alimentare, de unghiul α și de viteza de rotație. Relațiile obținute cu toate că sînt complicate permit calculul caracteristicilor mașinii fără a fi necesară încercarea experimentală, deci fie fără consum de energie fie în faza de proiectare. Trebuie ținut cont de faptul că s-au făcut și unele aproximări și ca atare expresiile deduse nu sînt întru totul exacte. Pentru studiul precis al caracteristicilor mașinii, sau cu aproximații admise în inginerie, se poate apela la integrarea sistemului de ecuații diferențiale ce descrie funcționarea mașinii printr-o metodă numerică, calculele efectuîndu-se cu un calculator cifric [5].

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THE EQUATIONS OF THE HALF WAVE CONTROL OF THE COMMUTATORLESS REPULSION MOTOR

(Summary)

The equations of the primary and secondary currents of a repulsion motor with SCR commutator (commutatorless) motor are derived. Also, the relationship of the electromagnetic torque in instantaneous and mean values are obtained for one half wave control.

The half wave control of commutatorless repulsion motor are used for small motors drives, the field in wich the approximations of calculus are valid.

As a consequence, the determination of the commutatorless repulsion motor performances is possible without a load test.

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CONTRIBUȚII LA MODELAREA SARCINILOR REȚELELOR ELECTRICE DE DISTRIBUȚIE URBANĂ

DE

GH. GEORGESCU, AL. POEATĂ și NINA POEATĂ

1. Introducere

Extinderea continuă a domeniilor de utilizare a energiei electrice la consumatorii urbani face ca prin rețelele de distribuție să se vehiculeze o cantitate de energie din ce în ce mai mare. Pe de altă parte, condițiile energetice actuale impun optimizarea consumurilor proprii tehnologice în rețele, motiv pentru care conducerea operativă și de perspectivă a rețelelor urbane constituie o problemă de actualitate. În literatura tehnică sînt prezentate modele matematice privind conducerea automată a acestor rețele.

Ceea ce este necesar în prezent, după părerea noastră, așa cum de altfel rezultă și din preocupările altor cercetători, pentru conducerea automată, eficientă, a diferitelor procese din rețelele electrice, este stabilirea unor indicatori sau caracteristici informaționale ale rețelelor electrice, care să caracterizeze complet starea rețelei, respectiv regimurile rețelei.

Totodată, se pune problema ca numărul datelor informaționale să fie redus la strictul necesar, iar obținerea lor să se facă cu minimum de cheltuieli.

Cele arătate se referă în special la rețelele de distribuție urbană, care au un număr foarte mare de noduri și consumatori, iar urmărirea individuală a regimurilor din fiecare nod este practic imposibilă. Factorul perturbator esențial, care condiționează deciziile în conducerea operativă și de perspectivă a unei rețele, îl reprezintă natura și dinamica sarcinii din nodurile rețelelor.

În vederea prezentării într-o formă cît mai concentrată a informațiilor privind sarcinile din nodurile rețelelor s-au impus metodele statistico-probabilistice. Recent, în literatura tehnică au fost publicate informații de acest gen, concretizate sub forma graficelor tip de sarcină, obținute prin prelucrarea statistică a unui număr mare de informații [1]. Precizăm că graficele de sarcină tip constituie „informația primară", cu ajutorul căreia se pot modela cu ușurință sarcinile din nodurile rețelelor, dacă se dispune de un minim de „informații secundare" din noduri.

Pe baza unui număr mare de măsurători efectuate de autori în cîteva rețele urbane, în decurs de patru ani, în scopul identificării principalilor indicatori care caracterizează sarcinile acestor rețele, publicate anterior [2]...[4], autorii și-au propus să stabilească graficele de sarcină tip, pentru rețelele urbane din țara noastră.

Măsurătorile au constat în ridicarea unui număr mare de grafice de sarcină activă și reactivă, pe perioada de vară și iarnă, în zilele lucrătoare și de repaus, în diferite puncte caracteristice ale rețelei și anume : barele de joasă tensiune ale posturilor de transformare, barele de medie tensiune ale punctelor de alimentare și stațiilor de transformare, firidele blocurilor de locuințe, pe un interval de 24 ore.

2. Stabilirea graficelor de sarcină tip

La stabilirea graficelor tip s-a plecat de la ipoteza că sarcinile din nodurile reţelelor de distribuţie urbană pentru diferite ore, „j", constituie mărimi aleatoare, care respectă legea de distribuţie normală. Pentru determinarea caracteristicilor probabile ale sarcinilor orare pe perioada unui an, folosind informaţii din diverse perioade ale anului şi în diferite puncte, legea de distribuţie a sarcinilor orare s-a scris sub forma :

$$\psi=\beta_1\psi_1+\beta_2\psi_2+\ldots+\beta_n\psi_n,\ \sum_1^n \beta_j=1, \tag{1}$$

încît valoarea medie, iregularitatea, dispersia şi abaterea standard se determină cu relaţiile :

$$\bar{C}_\psi=\int_{-\infty}^{\infty} C(\beta_1\psi_1+\beta_2\psi_2+\ldots+\beta_n\psi_n)\mathrm{d}C=\beta_1\bar{C}_1+\beta_2\bar{C}_2+\ldots+\beta_n\bar{C}_n, \tag{2}$$

$$(C^2)_\psi=\sum_{j=1}^n \beta_j(\bar{C}_j^2)=\sum_{j=1}^n \beta_j(1+k_{v,j}^2)(\bar{C}_j)^2, \tag{3}$$

$$D_\psi(C)=\sum_{j=1}^n \beta_j(\bar{C}_j^2)-\left(\sum_{j=1}^n \beta_j\bar{C}_j\right)^2. \tag{4}$$

$$\sigma_\psi(C)=\sqrt{\sum_{j=1}^n \beta_j(\bar{C}_j^2)-\left(\sum_{j=1}^n \beta_j\bar{C}_j\right)^2}. \tag{5}$$

În fig. 1...3 sînt prezentate graficele tip de sarcină activă, reactivă şi aparentă, obţinute pentru barele de joasă tensiune ale posturilor de transformare care deservesc consumatori casnici-edilitari din cartierele de locuinţe, precum şi dispersiile sarcinilor orare.

Valorile medii şi dispersiile sarcinilor orare indicate în graficele tip de sarcină sînt raportate la valoarea medie a graficului de sarcină.

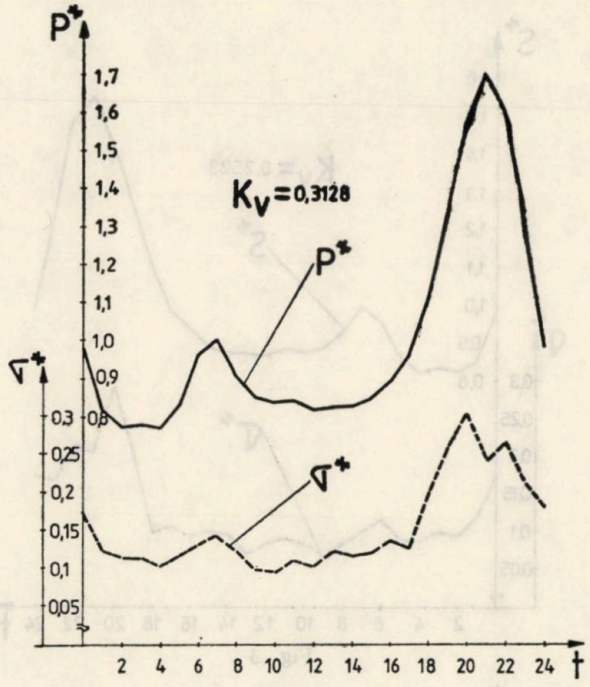
Unul din parametrii importanţi care caracterizează graficele tip de sarcină este coeficientul de variaţie, K_v, definit ca raportul dintre valoarea abaterii standard şi valoarea medie :

$$K_v=\frac{\sigma(C)}{\bar{C}}. \tag{6}$$

În cazul în care se consideră şi dispersiile valorilor orare ale graficului de sarcină, coeficientul de variaţie al sarcinii se determină cu relaţia :

$$K_v=\sqrt{\frac{1}{24}\sum_1^{24}(C_j^{*2}+\sigma_j^{*2})-1}, \tag{7}$$

unde : C_j^* — valoarea sarcinii la ora j din graficul tip de sarcină, σ_j^* — dispersia sarcinilor la ora j din graficul tip de sarcină, raportat la valoarea medie a sarcinii zilnice.


P*
1,7
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$K_V = 0{,}3128$
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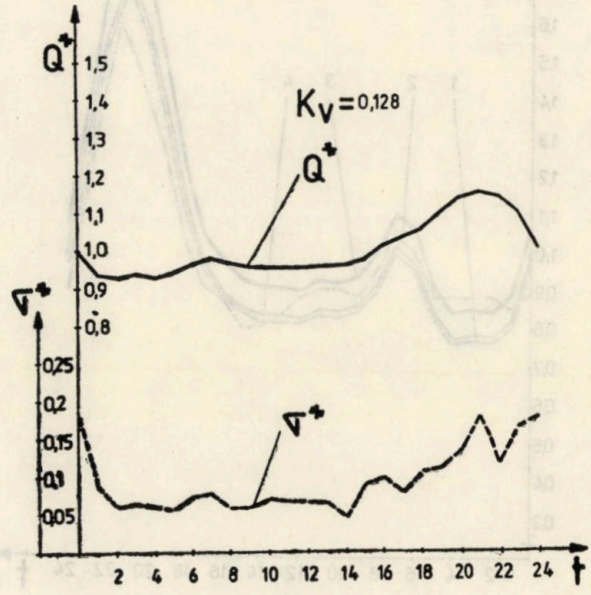
Fig. 1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$K_V = 0{,}128$
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σ*
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σ*
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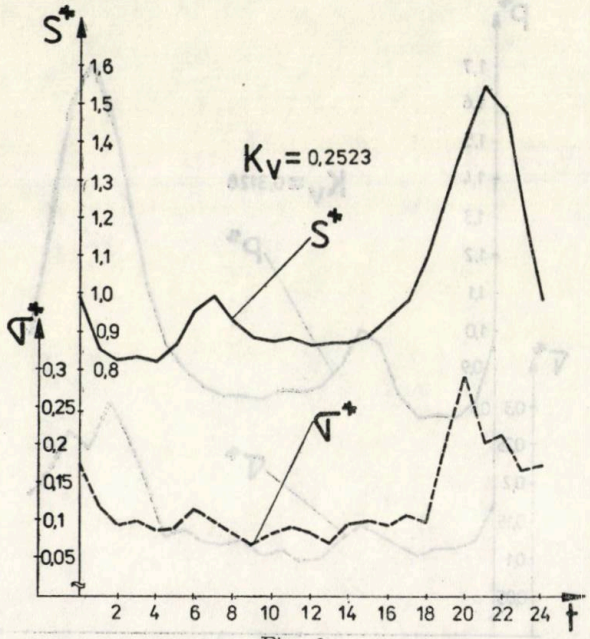
Fig. 2


S*
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Kv = 0,2523
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V*
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V*
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Fig. 3

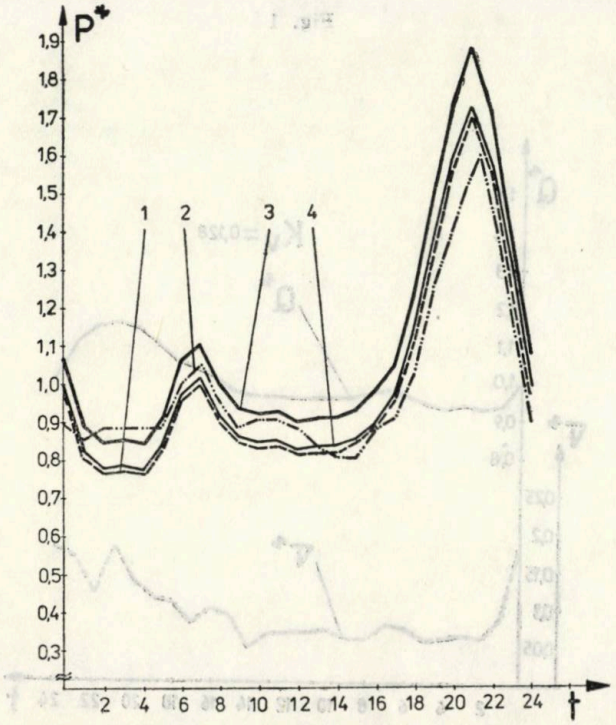
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Fig. 4

Un alt coeficient, care caracterizează graficele tip, este coeficientul de corelație dintre sarcinile a două grafice, determinat cu relația :

$$r(C_1, C_2) = \frac{(\overline{C_1 C_2}) - (\bar{C}_1 \bar{C}_2)}{\sigma(C_1)\sigma(C_2)} . \tag{8}$$

Cu ajutorul coeficienților de variație și corelație se poate stabili echivalența dintre graficele tip de sarcină pentru diverși consumatori. Dacă la două grafice se constată că coeficientul de corelație are o valoare apropiată de unitate, iar coeficienții de variație pentru cele două grafice au valori aproximativ egale, se poate afirma că graficele de sarcină sînt de același tip.

Pentru graficele tip de sarcină activă, reactivă și aparentă, prezentate în lucrare, ale posturilor de transformare urbane din țara noastră, s-au determinat coeficienții de variație, luînd în considerație și dispersiile sarcinilor orare, rezultatele fiind consemnate în fig. 1...3.

Modelarea graficelor de sarcină (activă, reactivă și aparentă) pentru un nod al rețelei, unde se cunoaște graficul tip al consumatorilor alimentați, se poate realiza prin asocierea la graficul tip, care constituie informația primară, a unei informații secundare, cum ar fi : sarcina la o oră oarecare din zi, sarcina maximă zilnică, energia consumată într-o anumită perioadă, coeficientul de încărcare a transformatorului.

2.1. Determinarea graficului de sarcină dintr-un nod al rețelei, prin asocierea la graficul tip al consumatorilor alimentați, a unei măsurători directe efectuate la o oră k, în nodul respectiv, ca informație secundară

În acest caz, sarcina, la ora j, a graficului se determină cu relația :

$$C_j = \frac{C_k C_j^*}{C_k^*} , \tag{9}$$

unde : C_k — sarcina măsurată la ora h, C_j^*, C_k^* — valorile raportate ale sarcinilor la ora j și respectiv k din graficul tip.

Pentru exemplificare, în fig. 4 sînt prezentate rezultatele obținute cu ajutorul relației (9), pentru un post de transformare din rețeaua de distribuție urbană, unde curba *1* reprezintă graficul tip de sarcină activă, curba *2* — graficul înregistrat al postului de transformare, curba *3* — graficul de sarcină obținut folosind ca informație secundară o măsurătoare la ora 7, curba *4* — graficul de sarcină obținut folosind ca informație secundară o măsurătoare la ora 13.

Analizînd curbele din fig. 4 se constată că graficul reprezentat de curba *4* are abateri mai mici față de graficul înregistrat (curba *2*), în comparație cu graficul redat de curba *3*. De aceea, în cazul în care se adoptă această metodă pentru determinarea graficelor de sarcină în nodurile rețelelor, recomandăm să nu se efectueze măsurători în perioadele cu variații mari de sarcină, adică în primele ore ale dimineții și seara, pentru consumatorii casnici-edilitari.

2.2. Determinarea graficului de sarcină dintr-un nod, prin asocierea la graficul tip a valorii maxime a sarcinii din nodul respectiv, ca informație secundară

Sarcina la ora j a graficului, în acest caz, se determină cu relația :

$$C_j = \frac{C_{\max} C_j^*}{C_{\max}}, \tag{10}$$

unde : $C_{\max}$ — sarcina maximă măsurată în nod, $C_{\max}^*$ — sarcina maximă din graficul tip, C_j^* — valoarea sarcinii la ora j din graficul tip.

În fig. 5 *a* și 5 *b* se reprezintă graficele de sarcină activă și reactivă obținute cu relația (10), pentru un post de transformare (curbele *3*) curbele *1* reprezentînd graficele tip de sarcină activă și respectiv reactivă, iar curbele *2* — graficele de sarcină activă și reactivă, înregistrate în postul de transformare.

2.3. Determinarea graficului de sarcină dintr-un nod al rețelei, cunoscînd graficul tip al consumatorilor și necesarul de energie dintr-o anumită perioadă în nodul respectiv, ca informație secundară

În acest caz, sarcina la ora j a graficului se determină cu relațiile :

$$C_j = \frac{W_{zi}}{24} C_j^*, \tag{11}$$

$$C_j = \frac{W_T}{24 N_T} C_j^*, \tag{12}$$

unde : W_{zi} — energia electrică necesară într-o zi (24 ore), W_T — energia electrică necesară într-o perioadă T, N_T — numărul de zile din perioada T, C_j^* — valoarea sarcinii la ora j din graficul tip.

2.4. Determinarea graficului de sarcină dintr-un nod al rețelei, prin asocierea la graficul tip de sarcină a coeficientului de încărcare a transformatorului din nod la sarcina medie sau la sarcina maximă, ca informație secundară

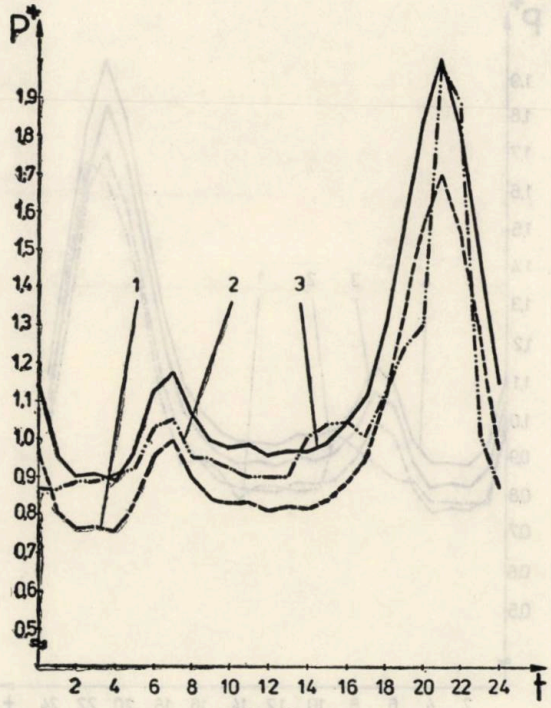
Sarcina, la ora j, a graficului se determină, în acest caz, cu relațiile

$$C_j = \overline{K}_{mc} S_n C_j^*, \tag{13}$$

$$C_j = \frac{\overline{K}_{Mc} S_n}{C_{\max}^*} C_j^*, \tag{14}$$

unde : $\overline{K}_{mc}$ — coeficientul mediu statistic de încărcare a transformatorului la sarcina medie, $\overline{K}_{Mc}$ — coeficientul mediu statistic de încărcare a transformatorului la sarcina maximă, S_n — puterea nominală a transformatorului, C_j^* — valoarea sarcinii, la ora j, din graficul tip, $C_{\max}$ — valoarea sarcinii maxime din graficul tip.

Pe baza graficelor zilnice de sarcină activă și reactivă, ridicate de pe barele de joasă tensiune ale posturilor de transformare care deservesc consumatori casnici-edilitari, în diverse perioade ale anului, s-au stabilit valorile medii și dispersiile pentru coeficienții de încărcare a transformatoarelor din posturile de transformatoare urbane, la sarcina medie și la sarcina maximă activă, reactivă, și aparentă, în regim de iarnă și regim de vară. Valorile acestor coeficienți sînt prezentate în tabelul 1.


P*
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t

Fig. 5 *a*

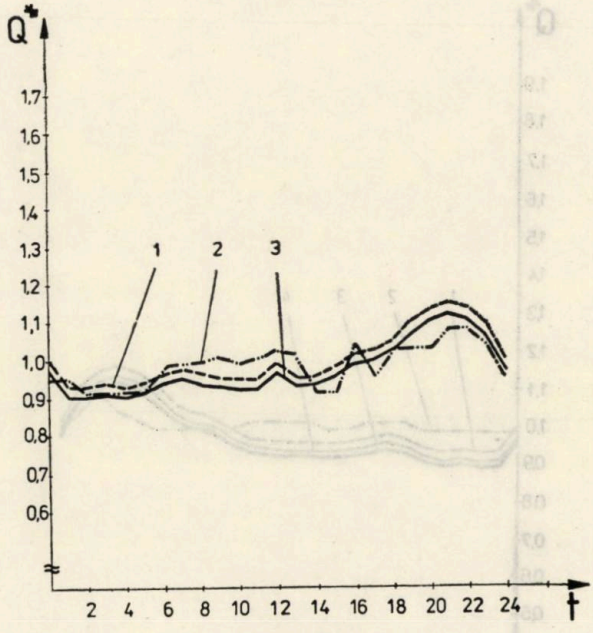
Q*
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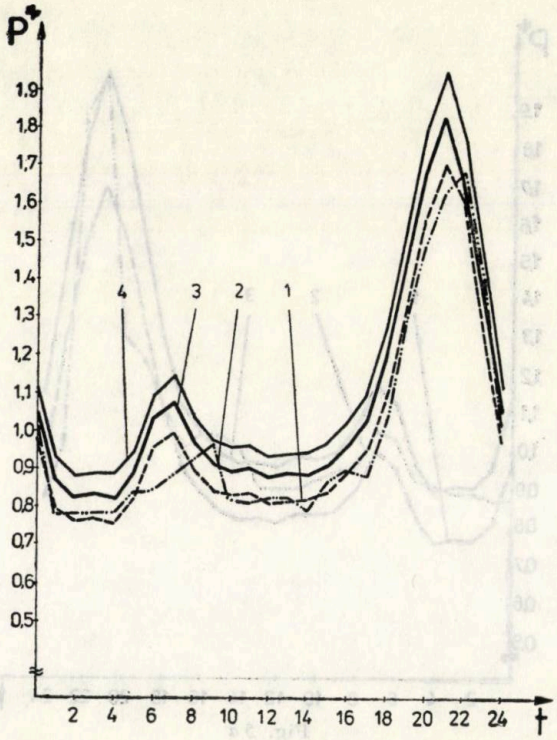
Fig. 5 *b*


P+
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4
3
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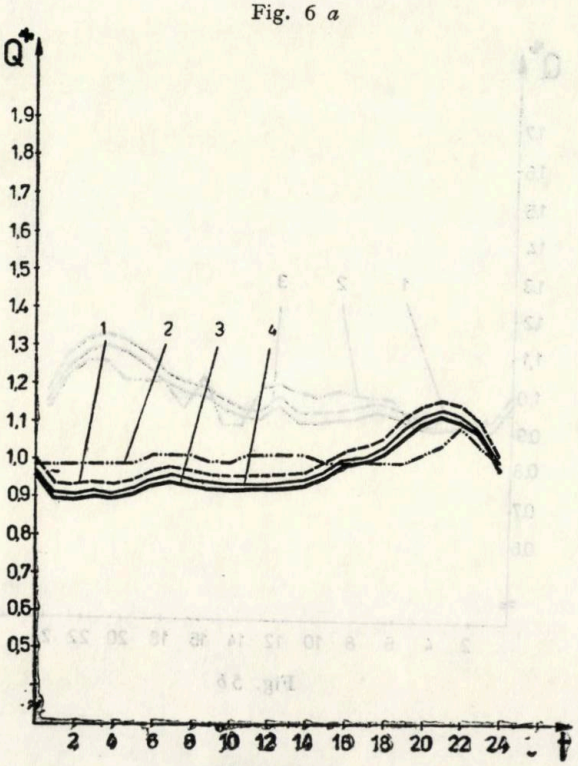
2 4 6 8 10 12 14 16 18 20 22 24 t

Fig. 6 *a*


Q+
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0,8
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0,6
0,5
1
2
3
4
2 4 6 8 10 12 14 16 18 20 22 24 t

Fig. 6 *b*

Tabelul 1

Sarcina		K			
		Regim de vară		Regim de iarnă	
		val. medie	dispersia	val. medie	dispersia
activă	medie	0,181	0,0489	0,201	0,047
	maximă	0,326	0,0878	0,332	0,0868
reactivă	medie	0,120	0,0293	0,121	0,029
	maximă	0,141	0,0471	0,142	0,046
aparentă	medie	0,218	0,0598	0,235	0,0575
	maximă	0,352	0,09162	0,359	0,0906

În fig. 6 *a* şi 6 *b* sînt reprezentate graficele de sarcină activă şi reactivă obţinute pentru un post de transformare, cu ajutorul relaţiilor (13) şi (14) (curbele *3* şi *4*), curbele *1* şi *2* reprezentînd graficele tip şi respectiv graficele înregistrate ale postului de transformare.

Tot din prelucrarea statistică a graficelor de sarcină înregistrate pe barele de joasă tensiune ale posturilor de transformare urbane s-au stabilit valorile medii şi dispersiile, pentru factorii de putere ai sarcinilor orare. Rezultatele obţinute sînt reprezentate în fig. 7.

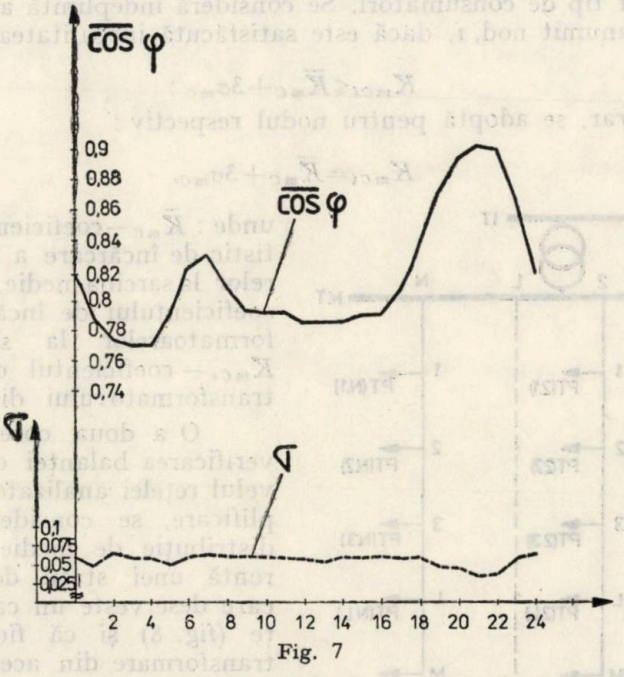

Fig. 7

Cunoscînd valorile cos φ_j, graficele de sarcină activă şi reactivă pentru un post de transformare se pot determina şi cu ajutorul relaţiilor :

(15) $$P_j=\bar{K}_{ms}S_nS_j^*\,\overline{\cos}\,\varphi_j,\quad Q_j=\bar{K}_{ms}S_nS_j^*\,\overline{\sin}\,\varphi_j,$$

(16) $$P_j=\frac{\bar{K}_{MS}S_n}{S_{\max}}S_j^*\,\overline{\cos}\,\varphi_j,\ Q_j=\frac{\bar{K}_{MS}S_n}{S_{\max}^*}S_j^*\,\overline{\sin}\,\varphi_j,$$

(17) $$P_j=\frac{\sqrt{3}UI_k\,\overline{\cos}\,\varphi_k}{P_k^*}P_j^*,\ Q_j=\frac{\sqrt{3}UI_k\,\overline{\sin}\,\varphi_k}{Q_k^*}Q_j,$$

unde: $\bar{K}_{ms}$, $\bar{K}_{Ms}$ — coeficientul mediu statistic de încărcare a transformatoarelor la sarcina medie şi respectiv maximă aparentă, S_n — puterea nominală a transformatorului, $S_{\max}^*$, S_j^* — valoarea sarcinii aparente maxime şi respectiv la ora j, din graficul tip, $\overline{\cos}\,\varphi_j$ — factorul de putere mediu statistic la ora j, I_k — curentul măsurat la ora k, P_j^*, Q_j^*, P_k^*, Q_k^* — sarcina activă şi reactivă din graficele tip, la ora j şi respectiv k.

3. Corecţii

În vederea creşterii preciziei modelării graficelor de sarcină din nodurile reţelelor de distribuţie urbană prin metodele expuse, sînt necesare, în principal, două corecţii [1].

O primă corecţie, pentru graficele de sarcină obţinute cu relaţiile (9)... (12), (17), se aplică în funcţie de abaterea coeficientului de încărcare a transformatoarelor, din nodurile analizate, de la valoarea medie determinată pentru acest tip de consumatori. Se consideră îndeplinită această condiţie pentru un anumit nod, i, dacă este satisfăcută inegalitatea:

$$K_{mCi}\leqslant\bar{K}_{mC}+3\sigma_{mC};$$

în caz contrar, se adoptă pentru nodul respectiv:

(18) $$K_{mCi}=\bar{K}_{mC}+3\sigma_{mC},$$

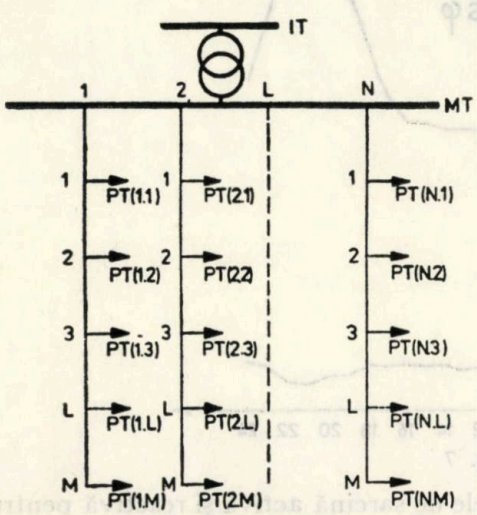

Fig. 8

unde: $\bar{K}_{mC}$ — coeficientul mediu statistic de încărcare a transformatoarelor la sarcina medie, σ_{mC} — dispersia coeficientului de încărcare a transformatoarelor la sarcina medie, K_{mCi} — coeficientul de încărcare a transformatorului din nodul i.

O a doua corecţie constă în verificarea balanţei de puteri la nivelul reţelei analizate. Pentru exemplificare, se consideră reţeaua de distribuţie de medie tensiune aferentă unei staţii de transformare care deserveşte un cartier de locuinţe (fig. 8) şi că fiecărui post de transformare din această reţea i s-a alocat un grafic de sarcină, după una din metodele expuse. Pentru a se rea-

liza balanța de puteri, trebuie ca pentru fiecare oră j din grafic să fie îndeplinită condiția :

$$\gamma C_{STj} - \sum_{i=1}^{n} \sum_{k=1}^{m} C_{i,k,j} = 0, \tag{19}$$

unde : γ — coeficient care ține seama de pierderile pe distribuitorii de medie tensiune, C_{STj} — sarcina vehiculată prin stație la ora j, $C_{i,k,j}$ — sarcina în nodul (i, k) la ora j, determinată prin una din metodele expuse. Această condiție, de regulă, nu se realizează direct și sînt necesare corecții :

$$\gamma C_{STj} - \sum_{i=1}^{n} \sum_{j=1}^{m} (C_{i,k,j} + \Delta C_{i,k,j}) = 0, \tag{20}$$

unde : $\Delta C_{i,k,j}$ reprezintă corecția sarcinii în nodul (i, k) la ora j. Aceste corecții pot fi determinate cu relația [1] :

$$\Delta C_{i,k,j} = \frac{\left(\gamma C_{STj} - \sum_{i=1}^{n} \sum_{j=1}^{m} C_{i,k,j}\right) C_{i,k,j}(1 - W_{i,k})}{\sum_{i=1}^{n} \sum_{j=1}^{m} C_{i,k,j}(1 - W_{i,k})}, \tag{21}$$

unde W_{ik} este coeficientul de încredere al informației secundare din nodul (i, k).

4. Program de calcul

Pe baza metodelor de modelare a sarcinilor și a rezultatelor obținute în urma prelucrării statistice a graficelor de sarcină, înregistrate în posturile de transformare care deservesc consumatori urbani, autorii au întocmit un program de calcul AL∅C-1, pentru modelarea, cu ajutorul calculatorului, a graficelor de sarcină activă, reactivă și aparentă din nodurile rețelelor de distribuție urbană de medie tensiune. Menționăm că programul AL∅C-1 permite analiza unei zone cu pînă la 200 de noduri, cu un timp de calcul redus.

5. Concluzii

1. Se prezintă graficele tip de sarcină activă, reactivă și aparentă de pe barele de joasă tensiune ale posturilor de transformare urbane din țara noastră, grafice care constituie informația primară necesară modelării sarcinilor în nodurile rețelelor electrice de distribuție urbană.

2. Se prezintă, sub formă de valori medii și dispersii, factorii de putere ai sarcinilor orare și coeficienții de încărcare a transformatoarelor din posturile de transformare urbane (fig. 7, tab. 1), valori care pot constitui informația secundară necesară modelării graficelor de sarcină, în cazul cînd nu se dispune de măsurători directe în rețea.

3. În cazul în care se folosesc, la modelarea graficelor de sarcină, ca informații secundare, măsurători directe, se recomandă ca acestea să nu se efectueze în perioadele cu variații mari de sarcină, adică în primele ore ale dimineții (5—8) și seara (18—23), pentru consumatorii casnici-edilitari.

4. Din prelucrările efectuate pe grafice de sarcină înregistrate în ani diferiți (1977—1980) s-a constatat că graficele tip pentru consumatorii casnici-edilitari variază puțin de la un an la altul, reactualizarea lor nefiind necesară anual, ci odată la 4—5 ani.

5. Programul AL⌀C-1, întocmit pentru modelarea sarcinilor conform metodelor și corecțiilor expuse, poate fi folosit ca subprogram în cadrul programelor elaborate în vederea adoptării deciziilor privind funcționarea optimă a rețelelor de distribuție urbană (stabilirea nivelului optim de compensare a sarcinilor reactive din posturile de transformare, debuclarea rețelei de medie tensiune, reducerea consumurilor proprii tehnologice, stabilirea nivelului optim de tensiune etc.).

Primită la 8 aprilie 1982

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Catedra de electroenergetică
și
Catedra de electronică și calculatoare

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CONTRIBUTIONS AU MODÉLAGE DES CHARGES DES RÉSEAUX ÉLECTRIQUES URBAINS

(Résumé)

L'étude statistique d'un grand nombre d'informations concernant les charges des réseaux électriques urbains a conduit aux courbes typiques de charge, données par les valeurs moyennes et les variances horaires.

Ces courbes ont été utilisées au modélage des charges attachées aux noeuds d'un réseau urbain. On présente aussi un programme de calcul élaboré à ce propos.

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L'INFLUENCE DES PARAMÈTRES DU RÉSEAU ÉLECTRIQUE SUR LES GRANDEURS STATISTIQUES DES VARIATIONS DE LA TENSION

PAR

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1. Introduction

Les tension des réseaux sont depuis longtemps dans une liste limitative de valeurs assez stables dans le temps, pour que les constructeurs dimensionnent leurs appareils au mieux. Le fonctionnement de ces appareils peut être considéré comme satisfaisant tant que la tension sous laquelle ils sont alimentés reste dans une certaine plage autour de sa valeur nominale. En pratique, même à l'intérieur de ces plages, le service rendu par les appareils varie en fonction de la tension et du temps d'alimentation à cette tension [1].

Pour considérer les variations de la tension dans le temps on utilise les grandeurs statistiques, à savoir [1] :

a) l'écart relatif de la tension, dans un point donné (j) et à un instant (i), donné :

$$u_{ji} = \frac{U_{ji} - U_n}{U_n} 100, [\%] ; \tag{1}$$

b) la valeur moyenne d'une série d'écarts relatifs de tension dans un point (j) :

$$u_j = \sum_{i=1}^{24} u_{ji}, [\%] ; \tag{2}$$

c) l'écart type d'une série d'écarts relatifs de tension :

$$\sigma_{uj} = \sqrt{\frac{1}{24} \sum_{i=1}^{24} (u_{ji} - \overline{u}_j)^2} = \overline{u_j^2} - \overline{u_j}^2. \tag{3}$$

Dans les relations ci-dessus U_n représente la tension nominale du réseau électrique.

On sait que la tension dans un point (j) d'un réseau est parfaitement caractérisée si l'on connaît sa valeur moyenne, $\overline{u_j}$ et son écart type, σ_{uj}.

Le réglage de la tension dans un réseau électrique consiste à y maintenir, à tout instant, une tension aussi proche que possible de la valeur nominale, U_n, caractérisant ce réseau. Ce résultat est obtenu lorsque simultanément la valeur moyenne, u_j, des écarts relatifs de tension et l'écart type, σ_{uj}, pour le point représentatif du réseau sont si réduits que possible.

Le réglage de la tension dans les réseaux électriques a évolué en même temps que le développement de ceux-ci. L'époque des réseaux radiaux est dépassée et dans les réseaux modernes, compte tenu qu'il y a différentes tensions superposées, il faut considérer aussi : le fonctionnement économique de l'ensemble du système énergétique, la minimisation des pertes de transport d'énergie, l'optimisation de la répartition des sources actives et réactives etc.

Dans ce qui suit on analyse l'influence des paramètres du réseau sur les grandeurs statistiques des variations de la tension dans le but de la conduite automatique des postes électriques.

2. La conduite automatique des postes électriques

La conduite du réseau électrique a connu un dévelopement rapide durant ces dernières années. Elle se caractérise par l'organisation régionale, le pays étant divisé en plusieurs régions opérationnelles, qui ont la responsabilité de l'exploitation des groupes de taille réduite, ainsi que des postes. Dans la phase de planification préliminaire le rôle du traitement des données dans ce système de conduite est très important.

Les progrès technologique dans le domaine des processeurs déterminent une évolution dans la conception des équipements des postes du réseau. Dans plusieurs pays cette évolution se manifeste par l'installation de systèmes programmées qui sont destinés à remplir des fonctions de conduite à distance, de conduite locale et d'automatisme [2].

Le but du réglage des transformateurs de distribution est de maintenir le niveau de tension chez le consommateur. Lors du fonctionnement en parallèle de plusieurs transformateurs, les réglages individuels doivent être choisis de façon à minimiser la circulation de la puissance réactive, tout en gardant le niveau de tension désiré. Une telle réalisation est relativement simple pour un système à deux transformateurs en parallèle mais, dès que l'exploitation normale d'un poste HT se base sur la mise en parallèle de plus de deux transformateurs, le système de réglage de la tension devient plus compliqué.

Dans [2] on prouve que de tous les réglages, le réglage „maître—esclave" représente la méthode la plus simple.

3. L'influence des paramètres du réseau

Pour le réglage optimal de la tension d'un réseau suivant le processus statistique il est nécessaire de connaître la manière dont les paramètres du réseau peuvent influencer les grandeurs statistiques. Dans ce sens on considère un réseau radial (Fig. 1) divisé en deux régions. Les transformateurs du réseau sont prévus avec les possibilités de réglage : $0, +5\%, -5\%$. Les caractéristiques ci-dessous sont trassées au niveau du jeu de barres à basse tension (BT) pour le réglage des transformateurs de la première région (I) et, séparément, de la deuxième région (II).

Dans les Fig. 2 et 3 est représentée la variation de la moyenne quadratique $I=\overline{u_j^2}$ en fonction du courant total d'alimentation du réseau (CAL) pour la prise moyenne, I_{BT}, et pour les réglages $\pm 5\%$, dans les régions I et II. On observe que la moyenne quadratique I augmente avec le courant d'alimentation et le meilleur de ces réglages est -5% pour la deuxième région.

Les courants des transformateurs de réseau sont considérés proportionnels à leurs puissance. Les calculs ont été effectués par voie automatique à l'aide du calculateur FELIX C-256, avec le programme RAM (Bibliothéque

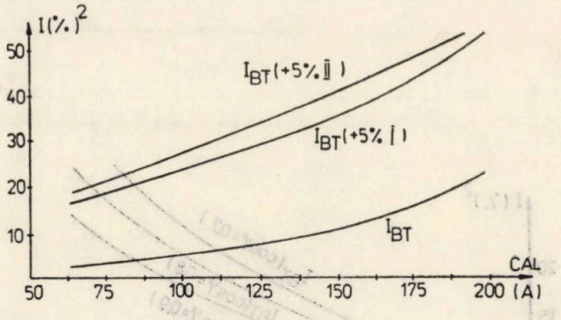

Fig. 2

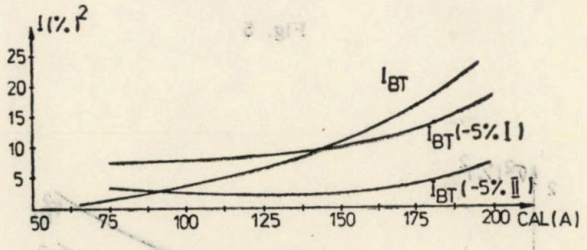

Fig. 3

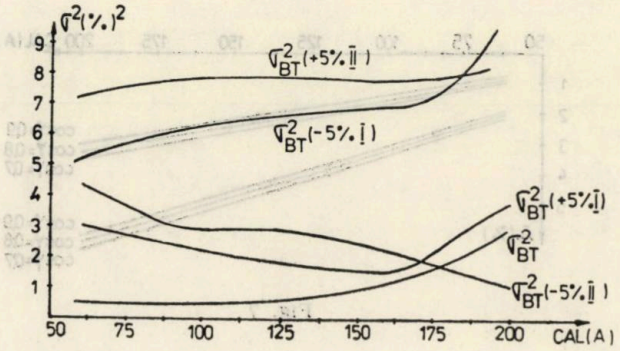

Fig. 4

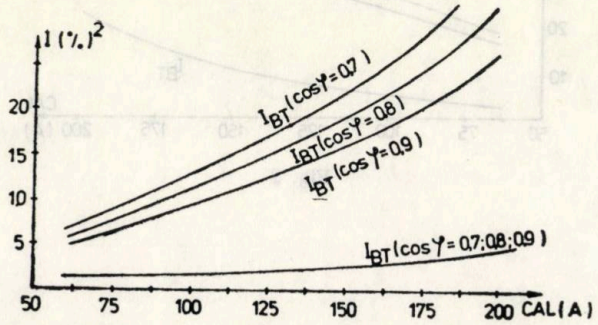

Fig. 5

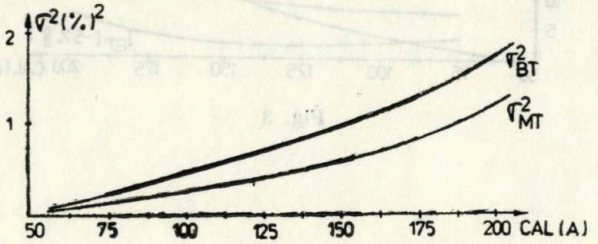

Fig. 6

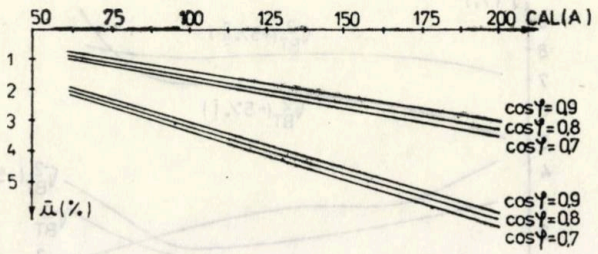

Fig. 7

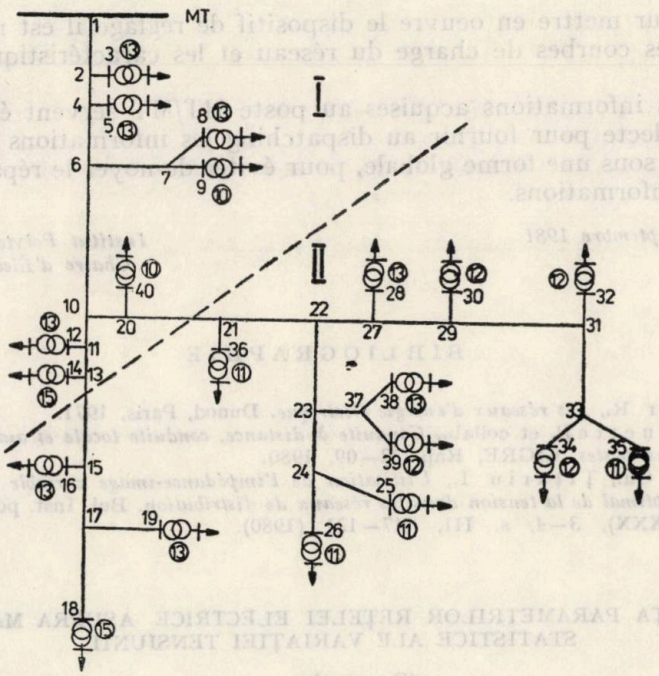

Fig. 1

ELECTRA) et un sous-programme spécial, pour la détermination des grandeurs statistiques.

Dans la Fig. 4 est représentée la modification de la variance en fonction du courant d'alimentation du réseau, σ^2(CAL). Les valeurs les plus grandes de la variance sont obtenues pour les réglages +5%(II) et −5%(I). On remarque le caractère nonlinéaire de ces caractéristiques.

Dans la Fig. 5 on observe que la moyenne quadratique I est plus grande pour la basse tension et qu'elle augmente avec la diminution du facteur de puissance ($\cos\varphi$). Pour le jeu de barres à moyenne tension, I ne dépend pas de la valeur du $\cos\varphi$.

La variance augmente d'une manière très forte avec le courant CAL et elle est beaucoup plus grande pour le jeu de barres à BT. La différence entre le jeu de barres à BT et à MT augmente, de même, avec le courant d'alimentation du réseau (Fig. 6). La différence entre la valeur moyenne à BT et à MT est grande, mais elle reste sans significations pour la variation de $\cos\varphi$ (Fig. 7).

Les fonctions σ^2(CAL) et I(CAL) donnent une image très exacte sur le processus de réglage de la tension dans un réseau électrique.

4. Conclusions

a) Les grandeurs statistiques peuvent servir pour la conduite automatique du réseau électrique.

b) Pour mettre en oeuvre le dispositif de réglage il est nécessaire de connaître les courbes de charge du réseau et les caractéristiques I (CAL), σ^2(CAL).

c) Ces informations acquises au poste HT/MT servent également de base de collecte pour fournir au dispatching les informations qui lui sont nécessaires, sous une forme globale, pour éviter de noyer le répartiteur sous un flot d'informations.

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INFLUENȚA PARAMETRILOR REȚELEI ELECTRICE ASUPRA MĂRIMILOR STATISTICE ALE VARIAȚIEI TENSIUNII

(Rezumat)

În scopul conducerii automate a rețelelor se studiază influența parametrilor acestora asupra mărimilor statistice care intervin în procesul de reglaj.

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LES EXPRESSIONS ANALYTIQUES DES INDICATIONS D'UN WATTMÈTRE (COMPTEUR) MONOPHASÉ MONTÉ DANS UN CIRCUIT TRIPHASÉ AVEC QUATRE CONDUCTEURS, EN RÉGIME DÉSÉQUILIBRÉ DE COURANTS

PAR

LAURIAN PREDA

1. Dans un système triphasé avec quatre conducteurs un wattmètre (compteur) monophasé peut être monté de neuf manières distinctes. On obtient, pour chaque montage, une déviation de l'aiguille indicatrice du wattmètre monophasé qui est notée par $a,b,\ldots,i$ (v. tableau 1). Les tensions étant symétriques, entre les neuf indications il existe les relations : $a+d+g=0$; $b+e+h=0$; $c+f+i=0$.

Tableau 1

	$\underline{I}_R$	$\underline{I}_S$	$\underline{I}_T$
$\underline{U}_R$	a	b	c
$\underline{U}_S$	d	e	f
$\underline{U}_T$	g	h	i

On présente les diagrammes des phases des tensions équilibrées et des composantes symétriques des courants déséquilibrés dans les six cas possibles (A, B, C, D, E, F)[1] de la répartition des trois mêmes valeurs effectives quelconques ($I_1=4$A, $I_2=3$A, $I_3=2$A) des courants déséquilibrés par rapport aux phases (Fig. 1 ... 6). De ces diagrammes phasoriels ils résultent 54 expressions analytiques des indications d'un wattmètre monophasé monté dans un système triphasé avec quatre conducteurs, pour les neuf manières distinctes de montage (Tableau 1) et les six cas possibles $(A, B, \ldots, F)$ de charge des phases par les mêmes courants (I_1, I_2, I_3).

2. Considérons le cas A $(I_R>I_S>I_T)$ et le diagramme phasoriel (Fig. 1) où l'on a noté par : $U_R=U_S=U_T=U$ la valeur effective des tensions de phase ; $I_1>I_2>I_3$ les valeurs effectives des courants de phase, quelqonques ; I_d, I_i, I_h — les valeurs effectives des composantes symétriques ; φ' — l'angle entre le phaseur de la tension, $\underline{U}$, et celui du courant $\underline{I}_d$ de la même phase ; φ'' — l'angle entre le courant $\underline{I}_i$ de la phase sur laquelle le courant est le plus fort et la tension de la même phase (dans ce cas l'angle entre $\underline{U}_R$ et $\underline{I}_{iR}$) ; φ''' — l'angle entre $\underline{I}_h$ et $\underline{U}$ de la phase sur laquelle le courant est le plus fort ; φ_1, φ_2, φ_3 — les déphasages entre les courants et les tensions correspondantes de la même phase. En choisissant comme origine de phase la tension $\underline{U}_R$, il résulte :

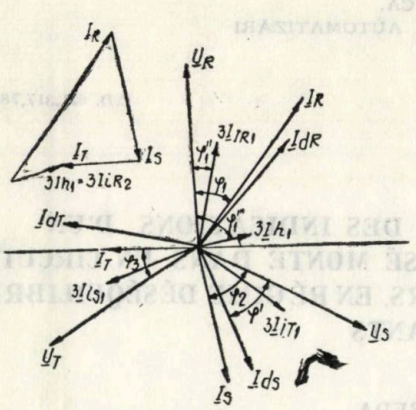

Fig. 1.—Cas A : $I_A > I_S > I_T$; $I_R = I_1 = 4$ A; $I_S = I_2 = 3$ A; $I_T = I_3 = 2$ A.

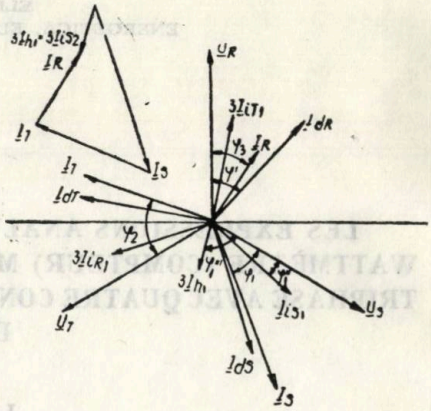

Fig. 2.—Cas B : $I_S > I_T > I_R$; $I_S = I_1 = 4$ A, $I_T = I_R = 3$ A, $I_R = I_3 = 2$ A

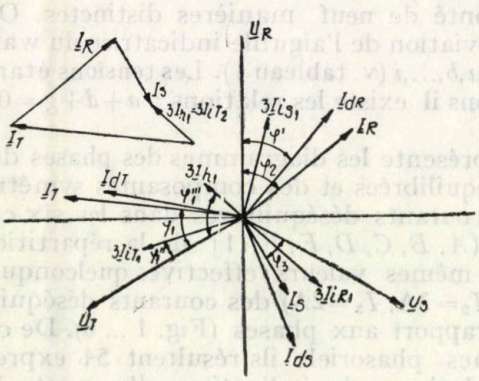

Fig. 3.—Cas C : $I_T > I_R > I_S$; $I_T = I_1 = 4$ A, $I_R = I_2 = 3$ A, $I_3 = I_2 = 2$ A.

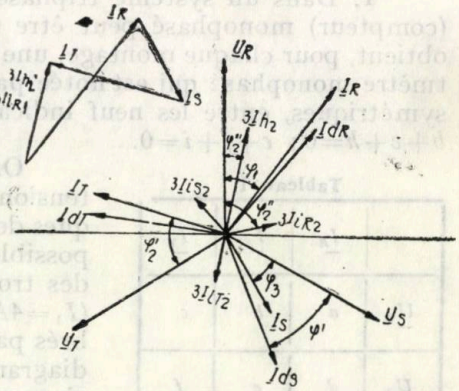

Fig. 4. — Cas D : $I_R > I_T > I_S$; $I_T = I_1 = 4$ A, $I_T = I_2 = 3$ A, $I_S = I_3 = 2$ A.

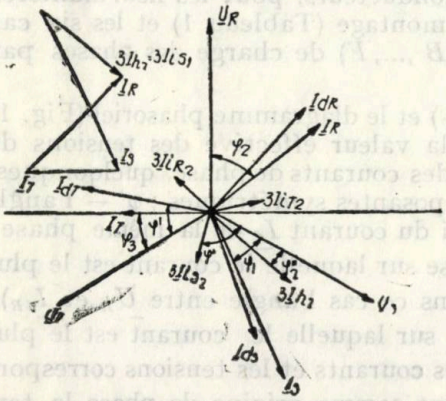

Fig. 5. — Cas E : $I_S > I_R > I_T$; $I_S = I_1 = 4$ A, $I_R = I_2 = 3$ A, $I_T = I_3 = 2$ A.

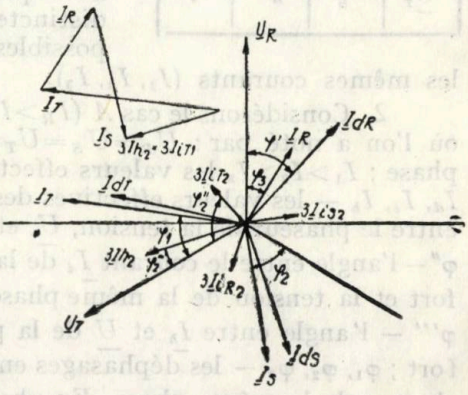

Fig. 6.— Cas F : $I_T > I_S > I_R$; . $I_T = I_1 = 4$ A, $I_S = I_2 = 3$ A, $I_S = I_3 = 2$A

$$(1)\quad \begin{cases} \underline{I}_R = I_1 e^{j\gamma_1}\ ;\ \underline{I}_S = I_2 e^{j(\varphi_2 - 2\pi/3)}\ ;\ \underline{I}_T = I_3 e^{j(\varphi_3 - 2\pi/3)}\ , \\ \underline{I}_{dRA} = \dfrac{1}{3}(I_1 e^{j\varphi_1} + I_2 e^{j\varphi_2} + I_3 e^{j\varphi_3}) = I_{dA} e^{j\varphi'_A} = I_d e^{j\varphi'}\ , \\ \underline{I}_{iRA} = \dfrac{1}{3}[I_1 e^{j\varphi_1} + I_2 e^{j(4\pi/3+\varphi_2)} + I_3 e^{j(2\pi/3+\varphi_3)}] = I_{iA} e^{j\varphi''_A} = I_{i1} e^{j\varphi''_1}\ , \\ \underline{I}_{hA} = \dfrac{1}{3}[I_1 e^{j\varphi_1} + I_2 e^{j(2\pi/3+\varphi_2)} + I_3 e^{j(4\pi/3+\varphi_3)}] = I_{hA} e^{j\varphi'''_A} = I_{h1} e^{j\varphi'''_1}\ . \end{cases}$$

Dans ce cas (Fig. 1), les déphasages entre les tensions et les courants sont donnés dans le tableau 2.

Tableau 2

	$\underline{I}_{dR}$	$\underline{I}_{iR}$	$\underline{I}_h$	$\underline{I}_{dS}$	$\underline{I}_{iS}$	$\underline{I}_h$	$\underline{I}_{dT}$	$\underline{I}_{iT}$	$\underline{I}_h$
$\underline{U}_R$	φ'	φ''_1	φ'''_1	$120°+\varphi'$	$240°+\varphi''_1$	φ'''_1	$240°+\varphi'$	$120°+\varphi''_1$	φ'''_1
$\underline{U}_S$	$240°+\varphi'$	$240°+\varphi''_1$	$240°+\varphi'''_1$	φ'	$120°+\varphi''_1$	$240°+\varphi'''_1$	$120°+\varphi'$	φ''_1	$240°+\varphi'''_1$
$\underline{U}_T$	$120°+\varphi'$	$120°+\varphi''_1$	$120°+\varphi'''_1$	$240°+\varphi'$	φ''_1	$120°+\varphi'''_1$	φ'	$240°+\varphi''_1$	$120°+\varphi'''_1$

En notant : $P = UI_d \cos\varphi'$, $Q = UI_d \sin\varphi'$, $p_1 = UI_{i1}\cos\varphi''_1$, $q_1 = UI_{i1}\sin\varphi''_1$, $p_{h1} = UI_{h1}\cos\varphi'''_1$, $q_{h1} = UI_{h1}\sin\varphi'''_1$, il résultent les neuf expresions analytiques des indications du wattmètre monophasé ci-dessous :

$$(2)\quad \begin{cases} a = P + p_1 + p_{h1},\quad b = -\dfrac{1}{2}P - \dfrac{\sqrt{3}}{2}Q - \dfrac{1}{2}p_1 + \dfrac{\sqrt{3}}{2}q_1 + p_{h1}, \\ c = -\dfrac{1}{2}P + \dfrac{\sqrt{3}}{2}Q - \dfrac{1}{2}p_1 - \dfrac{\sqrt{3}}{2}q_1 + p_{h1}, \\ d = -\dfrac{1}{2}P + \dfrac{\sqrt{3}}{2}Q - \dfrac{1}{2}p_1 + \dfrac{\sqrt{3}}{2}q_1 - \dfrac{1}{2}p_{h1} + \dfrac{\sqrt{3}}{2}q_{h1}, \\ e = P - \dfrac{1}{2}p_1 - \dfrac{\sqrt{3}}{2}q_1 - \dfrac{1}{2}p_{h1} + \dfrac{\sqrt{3}}{2}q_{h1}, \\ f = -\dfrac{1}{2}P - \dfrac{\sqrt{3}}{2}Q + p_1 - \dfrac{1}{2}p_{h1} + \dfrac{\sqrt{3}}{2}q_{h1}, \\ g = -\dfrac{1}{2}P - \dfrac{\sqrt{3}}{2}Q - \dfrac{1}{2}p_1 - \dfrac{\sqrt{3}}{2}q_1 - \dfrac{1}{2}p_{h1} - \dfrac{\sqrt{3}}{2}q_{h1}, \\ h = -\dfrac{1}{2}P + \dfrac{\sqrt{3}}{2}Q + p_1 - \dfrac{1}{2}p_{h1} - \dfrac{\sqrt{3}}{2}q_{h1},\quad i = P - \dfrac{1}{2}p_1 + \dfrac{\sqrt{3}}{2}q_1 - \dfrac{1}{2}p_{h1} - \dfrac{\sqrt{3}}{2}q_{h1}. \end{cases}$$

3. En gardant les mêmes valeurs effectives des tensions et des courants, mais en permutant d'une façon circulaire les courants par rapport aux phases (cas B et C des Fig. 2 et 3) ils résultent les neuf mêmes expressions analytiques des indications du wattmètre monophasé (2) mais pour d'autres situations $(a, b, \ldots, i)$ de montage de celui-ci (correspondant à la permutation circulaire). À la permutation circulaire des courants déséquilibrés par rapport aux phases (cas A, B, C et Fig. 1 ... 3), les composantes symétriques des courants déséquilibrés ne modifient pas leurs valeurs ; la direction de la composante directe reste la même, la composante inverse tourne dans le sens inverse avec $2\pi/3$ (cas B) respectivement $4\pi/3$ (cas C) et la composante homopolaire tourne en sens direct avec $2\pi/3$ (cas B), respectivement $4\pi/3$ (cas C).

Dans le cas où l'on maintient le même courant sur une phase et où l'on inverse les deux autres (par exemple le cas D, Fig. 4, par rapport au cas A, Fig. 1) les composantes inverses et les composantes homopolaires échangent entre elles leurs positions, la composante inverse devenant homopolaire et réciproquement (Fig. 1 et 4) ayant les déphasages $\varphi_2'' = \varphi_1'''$ respectivement $\varphi_2''' = \varphi_1''$. En notant : $p_2 = UI_{i2} \cos \varphi_2''$, $q_2 = UI_{i2} \sin \varphi_2''$, $p_{h2} = UI_{h2} \cos \varphi_2'''$, $q_{h2} = UI_{h2} \sin \varphi_2'''$, ils résultent les mêmes expressions analytiques des indications du wattmètre monophasé (2) tout comme dans le cas A dans lequel par contre tous les termes avec les indices 1 ont maintenant les indices 2. En permutant d'une façon circulaire (les cas E et F par rapport à D) apparaissent les mêmes modifications que les précédentes (les cas B et C par rapport à A).

4. De l'analyse des six cas (A, B, C, D, E, F) possib es, en écrivant les 54 expressions analytiques des indications d'un wattmètre monophasé monté dans un circuit triphasé avec quatre conducteurs, on constate, au déséquilibre de courants, que ces expressions analytiques peuvent être écrites sous la forme générale :

$$(3) \qquad (a, b, \ldots, i) = K_1 p + K_2 Q + K_3 p_1 + K_4 q_1 + K_5 p_{h1} + K_6 q_{h1} + K_7 p_2 + K_8 q_2 + K_9 p_{h2} + K_{10} q_{h2}.$$

Les valeurs des coefficients $K_1, K_2, \ldots, K_{10}$, pour chaque cas à part, sont indiquées dans le Tableau 3.

Dans le cas où l'on inverse les courants de deux phases (cas D, E, F par rapport à A, B, C), on trouve :

$$(4) \qquad I_{i1} = I_{h2}, \quad I_{h1} = I_{i2}, \quad \varphi_1'' = \varphi_2''', \quad \varphi_1''' = \varphi_2''.$$

Par conséquent les expressions analytiques (4) des indications du wattmètre monophasé dans les cas D, E, F peuvent être écrites en fonction seulement des grandeurs qui interviennent dans les cas A, B, C, compte tenu que :

$$(5) \qquad p_{i1} = p_{h2}, \quad p_{i2} = p_{h1}, \quad q_{i1} = q_{h2}, \quad q_{i2} = q_{h1}.$$

De (3), (5) et du Tableau 3, il résulte que les expressions analytiques des indications d'un wattmètre monophasé monté dans un circuit triphasé

Tableau 4

		$\underline{I}_R$						$\underline{I}_S$						$\underline{I}_T$					
		k_1	k_2	k'_3	k'_4	k'_5	k'_6	k_1	k_2	k'_3	k'_4	k'_5	k'_6	k_1	k_2	k'_3	k'_4	k'_5	k'_6
$\underline{U}_R$ a, b, c	A	1	0	1	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0
	B	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$
	C	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$
	D	1	0	1	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$
	E	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0
	F	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$
$\underline{U}_S$ d, e, f	A	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$
	B	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	1	0	1	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0
	C	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$
	D	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0
	E	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	1	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{3}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$
	F	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$
$\underline{U}_T$ g, h, i	A	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$
	B	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$
	C	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	1	0	1	0	1	0
	D	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$
	E	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$
	F	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	1	0	1	0

Tableau 3

		$\underline{I}_R$										$\underline{I}_S$										$\underline{I}_T$									
		k_1	k_2	k_3	k_4	k_5	k_6	k_7	k_8	k_9	k_{10}	k_1	k_2	k_3	k_4	k_5	k_6	k_7	k_8	k_9	k_{10}	k_1	k_2	k_3	k_4	k_5	k_6	k_7	k_8	k_9	k_{10}
$\underline{U}_R$ a, b, c	A	1	0	1	0	1	0	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	0	0	0	0
	B	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0
	C	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0
	D	1	0	0	0	0	0	1	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0
	E	1	0	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$
	F	1	0	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$
$\underline{U}_S$ d, e, f	A	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0
	B	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	0	0	0	0	1	0	1	0	1	0	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	0	0	0	0
	C	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0
	D	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$
	E	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	1	0	0	0	0	0	1	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0
	F	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$
$\underline{U}_T$ g, h, i	A	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0
	B	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0
	C	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	0	0	0	0	1	0	1	0	1	0	0	0	0	0
	D	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$
	E	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$
	F	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	1	0	$-\frac{1}{2}$	$\frac{\sqrt{3}}{2}$	0	0	0	0	$-\frac{1}{2}$	$-\frac{\sqrt{3}}{2}$	1	0	1	0	0	0	0	0	1	0	1	0

avec quatre conducteurs, dans le cas du déséquilibre de courants, peuvent être écrites seulement en fonction de six grandeurs, à savoir:

$$(6) \qquad (a, b, ..., i) = K_1P + K_2Q + K'_3p_1 + K'_4q_1 + K'_5p_{h1} + K'_6q_{h1}.$$

Les valeurs des coefficients sont données dans le Tableau 4.

Conclusions

1. Dans un circuit triphasé avec quatre conducteurs simples, symétriques, avec asymétrie de courants, un wattmètre monophasé, monté d'une façon quelconque, fournit une indication $a, b, ..., i$, proportionnelle aux puissances actives et réactives du circuit respectif (directes, grâce aux composantes directes des courants et d'asymétrie, grâce aux composantes inverses et homopolaires des courants) qui peuvent être exprimées analytiquement par une relation ayant dix termes de la forme (3) dont les coefficients sont conformément au Tableau 3.

2. Dans le cas où l'on effectue des permutations circulaires des trois mêmes courants déséquilibrés par rapport aux phases (cas A, B, C, respectivement D, E, F) les composantes symétriques ne modifient pas leurs valeurs et gardent le même déphasage par rapport à la tension de la phase sur laquelle le courant a la valeur maximum dans le cas considéré (Fig. 1 ... 6). Par conséquent les expressions analytiques des indications du wattmètre monophasé (v. (3) et Tableau 3) restent les neuf mêmes (v. (2)) mais pour d'autres situations $(a, b, ..., i)$ de montage de celui-ci (correspondant à la permutation circulaire).

3. Dans le cas où l'on maintien les trois mêmes courants déséquilibrés mais avec l'inversion de deux d'entre eux par rapport aux phases (cas D, E, F par rapport à A, B, C), les composantes directes restent les mêmes mais les composantes inverses et homopolaires échangent entre elles leurs positions, la composante inverse devenant homopolaire et réciproquement (v. Fig. 1...6 et les relations (4)).

4. Les expressions analytiques des indications du wattmètre monophasé peuvent être exprimées aussi par une relation (v. (6)) et Tableau 4) ayant seulement six termes.

5. Les expressions analytiques (2), (3), (6) permettent à trouver les montages [2], [3] qui mesurent sans erreur de méthode les puissances directes (actives et réactives) et celle de nonsymétrie (inverses—de disymétrie et homopolaire — d'asymétrie).

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EXPRESIILE ANALITICE ALE INDICAȚIILOR UNUI WATTMETRU (CONTOR) MONOFAZAT, MONTAT ÎNTR-UN CIRCUIT TRIFAZAT CU PATRU CONDUCTOARE, ÎN REGIM DEZECHILIBRAT DE CURENȚI

(Rezumat)

Utilizînd metoda componentelor simetrice se găsesc expresiile analitice (v. tabelele 3 și 4) ale indicațiilor unui wattmetru (contor) monofazat montat într-un circuit trifazat cu patru conductoare. Aceste expresii depind de modul în care wattmetrul (contorul) este conectat în circuit și de repartiția celor trei curenți dezechilibrați în raport cu cele trei faze.

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RECENZII

BOOK REVIEWS РЕЦЕНЗИЙ

GALLAWA R.L., *A User's Manual for Optical Waveguid eCommunications*. Information Gatekeepers Inc., Brooklyne, U.S.A., 1981, 287 pp.

Many monographs and handbooks were published last years in the rapidly developping field of optical communications. This one, an users oriented manual, is intended to fulfill the gap between specialists' and general public's literature. Indeed, it is adressed to the practicing communicator who is prospecting the technology, capabilities and profitability of optical waveguide digital communications. The subject is introduced by intuitive, less-mathematical possible notations, the most difficult mathematical topics being relegated to the appendixes.

The main topics developped by the author cover both the introductive problems in optical communications (Chapters 3 ... 5 : optical sources, optical detection, the transmission medium) and the practical aspects (Chapters 6 and 7 : modulation and demodulation, systems considerations and design). An Introduction and a 15-pages Glossary of therms help the newcomer to understand the book. In order to facilitate the reading, no references are given. However, an extensive bibliography, by subject matter, may be used by those who are interested in going deeper in the field.

If it is not difficult for a newcomer to read this manual, this does not means that it is lacking interest for other readers. They will find the book, which is written in a clear style, very attractive and a valuable reference or aide-mémoire.

H.-N. Teodorescu

TURNER L.W. (Edit.), *Electronics Engineer's Reference Book*. 4-th Ed., Newnes-Butterworths, London-Boston, 1976.

The book, a fourth edition almost entirely rewritten, represents the result of the efforts made by 62 specialists in various fields of Electronics. A wide variety of topics ranging from physical aspects and materials up to theory and applications of electronic devices and circuit s are covered in 27 sections : General Information, History of Electronics, General Physical Back - ground, General Background to Electromagnetic and Nuclear Radiation, Influence of Iono - sphere and Troposphere on Radio Wave Propagation, Electronic Materials and Components , Electron Valves and Tubes, Solid State Devices, Photo-electronic Devices, Electro-optic-Devices, Integrated Circuits, Microelectronics, Basic Electronics Circuits, Electronic Instrumentation and Measurements, Telecommunications, Sound and Video Recording, Electronic Music, Radar Systems, Electronics in Weather Forecasing, Radio Astronomy, Electronics in Space Explorations, Electronic Aids in Education, Public Address and Sound Reinforcemen t Systems, Electronics in Industry and Business, Electronic Applications in Transport, Par ticle Accelerators for Nuclear Research, Electronic Aids in Medicine.

The book represents a real reference one not only for electronic engineers but also for specialists in other fields interested in electronics principles and applications as it is very clearly written, principles, ideas and qualitative aspects being first of all emphasized. For further in - formations, reference lists are included at the end of each section.

L. Goraş

NORRIS B. (Edit.), *Power-transistor and TTL Integrated-circuit Applications*. Applicati ons Laboratory, Texas Instruments Ltd.

Written by engineering specialists in the Texas Instruments Applications Laboratory , the chapters in this volume have a broad industrial focus. Practical and application-oriented ,

they are of interest to a wide spectrum of engineers using semiconductor devices. Also, the book's coverage of inverter and audio amplifier design will interest engineers who are not specializing in the field.

The book is devided into four sections. The first covers power control with triacs, thyristors and unijunction circuits. Introductory chapters provide fundamentals, the construc tions, characteristics and comparisons made between triacs and thyristors, also examine thei[r] applications in basic circuits. Following chapters describe more specific applications such a[s] burst firing techniques and circuits and reversible d.c. supply. Extensive coverage is given to the theory, operation and circuits of unijunction transistors.

The second section, regarding the power transistor applications, concentrates on power supply designs, particularly of the switching mode type and inverters. A chapter is devoted to a TV application, which presents a regulator design procedure useful in all applications.

Section 3 is devoted to the design of audio amplifiers from 1 watt to over 100 watts. It also examines a high fidelity preamplifier and reveals how Super Silect transistors can provide considerably more power to audio amplifiers.

The final section shows how the most widely used family of circuits (TTL) can be applied. It gives basic descriptions and examples of how counters and registers may be used and connected and demonstrates a number of relatively simple circuit applications. The last chapter describes how TTL may be safely employed if its proposed environment is electrically noisy.

The material presented in this how-to-do-it guide represents original contributions by acknowledged experts in their specialities. Its proven principles and time-saving formulas, charts, tables and other practical data are directly applicable to work.

I. Bogdan

THELIEZ S., TOULOTTE J.-M., *GRAFCET et logique industrielle programée*. Eyrolles, Paris, 1980, VII + 137 pp.

L'ouvrage a pour but de présenter les méthodes de synthèse et de matérialisation des structures logiques industrielles programmables.

Dans le premier des cinq chapitres on examine les méthodes d'implantation des fonctions combinatoires sur le matériel programmé. Le deuxième chapitre présente les outils de description des automates séquenciels : le Grafcet et les réseaux Pétri. Dans les deux chapitres suivants sont présentées les méthodes d'implantation des outils de description, sans oublier les applications industrielles (dispositifs à logique câblé et les automates programmables). L'ouvrage se termine avec un chapitre bibliograhique.

Bien qu'il aie une extension relativement réduite, l'ouvrage constitue une aide réelle pour ceux qui désirent se familiariser avec les méthodes de la logique programmée.

C. Huţanu

JENSEN K., WIRTH N., *PASCAL. Manuel de l'utilisateur*. Eyrolles, Paris, 1981, 167 pp.

Le but de ce livre est de présenter un nouveau langage de programmation pour les ordinateurs numériques — Pascal, développé après 1968 et ayant comme base ALGOL 60.

L'ouvrage, destiné aux personnes qui sont déjà familiarisées avec la programmation, est constitué de deux parties : le manuel utilisateur et le Rapport Révisé.

La première partie est constituée de 14 chapitres, suivis par la bibliographie concernant le langage Pascal, six annexes et l'index des termes utilisés.

Les chapitres 0 à 12 (Introduction, Notations et vocabulaire, Le concept de donnée, L'en-tête de programme et la partie déclaration, Le concept d'action, Types scalaires et intervalles, Types structurés en général-tableau en particulier, Le type article (RECORD), Le type ensemble (SET), Les types fichiers (FILE), Types pointeurs, Procedures et fonctions, Entrées/Sorties) décrivent Pascal Standard.

Les chapitres 13 et 14 présentent la réalisation et l'utilisation de Pascal sur l'ordinateur Control Data 6 000. L'extension du langage, appelé Pascal 6 000—3.4, est utilisé sous le système d'exploitation SCOPE 3.4.

En annexes se trouvent rassemblées de nombreuses tables recapitulatives et les spécifications syntaxiques du langage Pascal.

Le Rapport, constitué de 14 chapitres, définit le Pascal Standard — qui représente la base commune aux diverses réalisations du langage — et peut servir de référence concise pour le programmeur ou pour le réalisateur de complilateur.

Le livre a un style didactique et utilise de nombreux exemples de programmes pour illustrer les particularités et les facilités du langage ; par conséquent l'ouvrage constitue un excellent guide pour ceux qui désirent apprendre Pascal.

N. Thirer

UNGER H.G., *Hochfrequenztechnik in Funk und Radar*. Verlag B.G. Teubner, Stuttgart, 1972, 224 S.

Dieses Lehrbuch führt den Ingenieur und Physiker in die wichtigsten hochfrequenztechnischen Grundlagen ein und berücksichtigt dabei insbesondere ihre Anwendungen in Rundfunk, Richtfunk, Satellitenfunk und Radar.

Vorausgesetzt werden beim Leser nur die Kenntnis der allgemeinen Grundlagen der Elektrotechnik und die einfachsten Begriffe aus der Nachrichtentechnik.

Der erste Kapitel, auch der umfangreichste, umfaßt die Problemen von Sende- und Empfangsantennen. Die theoretische Behandlung der Strahlung ist mit der Hilfe der linearen Antenne illustriert und die Kenngrößen von Antennen : Richtdiagramm und Antennengewinn, sind also eingeführt. Andere Kenngrößen wie Strahlungswiderstand, Eingangswiderst und Wirkfläche, Übertragungsfaktor sind anschaulich gemacht und daraus gewinnt man nützliche Beziehungen von praktischen Interesse.

Die wichtigsten Antennentypen wie : verlängerte und schwundmindernde Vertikalantennen, Rahmen- und Ferritantennen, Faltdipol und Breitbanddipole, Gruppenstrahler, Langdraht-, V- und Rhombusantennen, Wendelantennen, Flächen- und Schutzstrahler, sind vorgestellt.

Im zweiten Kapitel, *Wellenausbreitung*, mit einer mässigen mathematischen Behandlung und auf qualitativen Basis stellt man die Boden- und Raumwellen vor.

Der dritte Kapitel umfaßt die Funktion der Senderöhren: Sendetrioden, Strahltetroden, Klystronen, Wanderfeldröhren, Magnetronen und die wichtigsten Sendeschaltkreisen.

Die Hochfrequenz-Empfangsproblemen sind im vierten Kapitel behandelt. Eine spezielle Achtung ist dem Rauschen und der Empfindlichkeit gewidmet.

Der fünfte Kapitel, *Rundfunktechnik*, beschäftigt sich mit der Blockschaltdiagrammen der AM- und FM-Rundfunksendern, Hörrundfunkempfängern, Fernsehsendern und Fernsehempfängern.

Die Datenübertragungssystemen sind in der sechsten Kapitel, *Richtfunktechnik*, behandelt. Als Beispiel beschreibt man das Richtfunksystem FM1 800/6 000.

Der siebente Kapitel befaßt sich mit der Prinzipien der Satellitenfunk. Die erforderliche Sendeleistung der Bodenstation und der Satellitenstation sowie der erforderliche Empgangs-Signal/Rausch abstand und die Mitteln die das ermöglichen sind abgeschätzt .

Der achte Kapitel behandelt die Radartechnik. Einige Begriffe wie : Radarquerschnitt und Reichweite sind eingeführt, dann stellt man die Prinzipien der Impulsradaralagen vor.

Das Skriptum bildet Inhalt und Ergänzung einer Vorlesung, die die Studierenden der Elektrotechnik an der Technischen Universität, Braunschweig, Bundesrepublik Deutschland, in die Hochfrequenztechnik einführt. Es ist aber so ausführlich abgefaßt, das es sich auch zum Selbststudium und zur Einarbeitung eignet. Es wendet sich an die Studierenden der Elektrotechnik und Physik an Universitäten sowie an Elektroingenieure, Elektroniker und Physiker in der Praxis.

N. Alexandru

* * * *Optoelectronics Applications Manual*. The Applications Engineering Staff of the Hewlett-Packard Optoelectronics Division, Mc. Graw Hill Book Co., New York, 1977.

This book, printed in an attractive manner, is one of those monographs intended to fill a gap in the design engineers' literature. The four authors, Gage S., Evans D., Hodapp M. and Soreksen H., from HP Optoelectronic Division, are experienced fellows in the field covered by the book.

The manual treats about LED devices, including optoisolators, displays, LED to optical fiber connections and other related applications. The reader could find topics in a modern format on both theory and technology of the discrete LEDs, integrated circuits and circuits using LEDs.

A very attractive feature of this work is that various up to date applications are illustrating the theoretical topics.

The volume is devided into nine sections. The starting chapter is treating LED theory and is followed by a detailed chapter on LED lamps, describing their physical properties, electro-optical characteristics, packaging and applications in visualisation, lighting, signaling and communications.

Chapter 3 reviews the theory, characteristics, packaging and curent transfer ratio (CTR) degradation of the optoisolators and describes the analogue and digital applications of these devices. Techniques of design allowances for CTR degradation and applications are given.

The topic of displays, the wider application field of LEDs, is covered by two chapters, 5 and 6. Chapter 5 presents the various types of LED displays, display subsystems, driving conditions and circuits and the interfacing microprocessors systems to LED displays problems.

Chapter 6 is a good example of detailed applicative explanation of how appropriate use of technology can do a poor performance device to work with best performances. Numerous examples of techniques which can be used to improve display visualisation are given.

Measurement problems are discussed in Chapter 7, devoted to photometry and radiometry. Chapter 8 offers a brief review of the reliability of LED products and of test conditions. The book ends with a chapter dealing with mechanical handling, storage, soldering and socket mounting of LEDs.

The manual could be used by graduate students in electronics, research and design engineers and other people interested in the optoelectronics. It is a practical tool to those who whish to use the optoelectronic devices with a maximum efficiency and for those searching new applications.

H.-N. Teodorescu

COCQUEREZ J. P., DEVARS J., *Microprocesseurs à l'usage des électroniciens*. Ed. Eyrolles, Paris, 1980, VII + 163 pp.

L'ouvrage, ayant un caractère didactique prononcé, fait une présentation graduelle et bien équilibrée des problèmes émanant de l'utilisation des microprocesseurs. Le livre comprend neuf chapitres et cinq annexes.

Le premier chapitre traite les principaux éléments structuraux des machines à programme enregistré (mémoires, régistres, compteurs, UAL) de même que l'architecture des microprocesseurs. Les chapitres II at III sont destinés à la présentation des structures, du déroulement des instructions et des modes d'adressage des machines à programme enregistré. Les problèmes présentés dans les chapitres précédents sont repris dans le chapitre IV à travers des exemples concrets utilisant deux types de microprocesseurs : 6 800, de Motorola et 6 100, de Intersil.

Procédant de la même façon, les chapitres V et VI présentent les problèmes d'interfaçage des microsystèmes à microprocesseurs avec des exemplifications utilisant les deux types de microprocesseurs cités plus haut. Les chapitres suivants traitent l'organisation d'une carte micro-ordinateur (chap. VII) et la complémentarité qui existe entre la partie matérielle et la partie logique (chap. VIII). Le dernier chapitre traite les problèmes qui sont liés à la logique des microprocesseurs (programme monitor, les languages de programmation, programmes utilitaires, mémoires de masse).

Dans les annexes ont été traités les microprocesseurs utilisés dans les exemplifications, de même qu'un rappel sur les représentations octale at hexadécimale, le code ASCII et les éléments d'arithmétique binaire.

L'ouvrage s'adresse aux étudiants, mais aussi aux ingénieurs qui désirent connaître la technique des microprocesseurs.

C. Huţanu

MARSHALL G.J., *Principles of Digital Communications*. Mc Graw-Hill Book Ltd., London-N.Y., 1980, 170 pp.

The book is an introductory course to the modern area of digital communications. It is written in a very clear style, the principles being emphasized first of all. A reader who know nothing or very little about the subject can learn the fundamental ideas involved in digital communications. The main topics covered in the book are divided as follows : Communication systems and networks, Information theory (Coding and communications over noisy channels) and Digital transmission (Baseband pulse transmission, coded transmission and carrier-borne transmission). Each chapter ends with relative simple problems having solutions at the end of the book.

L. Goraş

* * * *Communications Equipment and Systems.* I.E.E. Communications, 1978.

This new book published by the Institution of Electrical Engineering is a valuable and very useful work.

It consists in 157 communications presented in the 10 sessions of the 1978 Conference which were grouped in three great themes : PTT Telecommunications, Civil Fixed and Mobile Radio Communications and Defence Communications. Every theme comprises very important communications covering domains of great interest, such as : Telephone Switching, Network Applications, Data Systems, Planning and Transmissions, PTARMIGAN System, etc.

The authors of the communications are very good specialists and work in the most famous companies from United Kingdom and outside (Marconi Communications Systems Ltd., Plessey Co. Ltd., Philips Research Laboratories, Harris Corporation, Racal-Milgo Ltd., Ceneral Electric Co. Ltd., Marconi Space and Defence Systems Ltd., etc.) as well as officers from the Ministry of Defence. They deal with microprocessor applications, computerised communications systems, modems, noise problems, etc. Almost all communications present blockdiagrams and even photos. Some of them (12) deal with the new NATO's project of communications system — PTARMIGAN.

This volume comprises also two overview addresses presenting general problems in Civil Fixed and Mobile Radio Communications and in Defence Communications.

I. Bogdan

* * * *Physics of Modern Materials* Lectures presented at an International Course at Trieste, from 29 March to 24 June 1978, organized by the International Centre for Theoretcal Physics, Trieste. Vol. **I**, 530 pp., Vol. **II**, 690 pp., IAEA, Vienna, 1980.

The annual Courses in Trieste, through scientific instructions their characters are interesting to follow so much by specialists and those who work in related field.

The lectures of March-June 1978 Session, with the general topic *Physics of Modern Materials*, are destined so much to physicists, engineers and technicians who work in electronics, electrotechnics, mechanics, etc.

The publication of these lectures, revised by the authors in 1979/1980, by IAEA Vienna, lay in the hands of those who wish to gather documentation about certain domains in physics of materials, chapters on general properties, mechanics, electronics, magnetics and optics of crystals and amorphous materials.

Presented by specialists from renowned universities centre or research centre from all over the world, the 28 lectures refer to topics of different interest : Characterization and growth, Defects, Mechanical properties, Metals and alloys, Semiconductors and devices, Amorphous and superionics.

Every chapter, after the presentation of basic elements, treat the principal aspects of every topic, indicating the modern directions of research, also concluding whith a selected bibliography. Some chapters also have list of widen bibliographies.

These two volumes impose so much, instruction, teaching, grafic achievement and their scientific content.

The second volume, presents in 80 pp. also the works *Topical Symposium on Classical Fluids* (6 communications) which took place on June 12— 13, 1978, whith the occasion of Prof. Ivar Waller (Upsala) birthday, author of important discoveries in modern physics.

Tr. Balan

BRIE C., GERBER R., *Les microprocesseurs en tranches. Technique et Documentation.* Paris, 1979, 246 pp.

Le livre, dont les auteurs sont des specialistes français bien connus dans le domaine l'automatisme et de l'informatique, contient une présentation détaillée des microprocesseurs de type „bit slice". Ces neufs chapitres renferment une présentation graduelle du problématique de ces dispositifs intégrés.

Le premier chapitre traite les technologies utilisées pour la fabrication des microprocesseurs, ensuite aborde le fonctionnement des circuits MSI/LSI qui se retrouvent à la base de la réalisation des microprocesseurs en tranches (multiplexeurs, régistres, mémoires, PLA, etc.). Les trois chapitres suivants présentent les problèmes généraux liés à la structure interne des micro- et miniordinateurs sans oublier les structures microprogrammables.

Dans les chapitres 5 et 6 sont présentés les principaux éléments des systèmes en tranches, avec des exemples : les séquenceurs Am 2 911 et Am 2 910 de AMD et unités de traitement. Les problèmes liés à la mise en fonction des systèmes à microprocesseurs en tranches son t traités dans le chapitre 7.

L'ouvrage se termine par la présentation de deux exemples dans deux chapitres : traitement des images (chap. 8) et l'application dans la téléinformatique (chap. 9).

Grâce à son mode judicieux de présentation et son dosage des problèmes étudiés l'ouvrage constitue un support réel pour ceux qui désirent approfondir leurs connaissances dans le domaine des microprocesseurs. *C. Huţanu*

MATTHEWS H. (Ed.), *Surface Wave Filters. Design, Construction and Use.* John Wiley a. Sons. New York-London-Sydney-Toronto, 1977, 521 pp.

Surface elastic waves have been studied intensively during the last two decades. The main idea is that an elastic surface mechanical wave that has velocities 10^4 times smaller than light is accompanied in a piezoelectric solid by an electromagnetic wave with the same small velocity and this wave can interact with the so-called interdigital electrodes that represent the transmitter and the receiver for these waves.

The book represents the result of the efforts made by 14 experts in this domain to cover a wide of subjects that represents the 11 chapters in which it is devided : Elastic Surface Waves, The Interdigital Transducer, Principles of Surface Wave Filter Design, Surface Wave Device Fabrication, Matching Networks and Packaging Structures, Surface Wave Bandpass Filters, Phase Code Generators and Correlators, Surface Wave Interdigital Electrode Cirp Filters, Reflection Grating Filters, Surface Wave Devices for Radar Equipment, Surface Wave Devices in Spread Spectrum Systems.

As it can be seen from the above, the initial chapters present principles and methods of design and construction while the next ones deal with properties and performances of filters. The last two chapters deals with the use of these filters in radar and communications. At the end of each chapter the reader will find much bibliographic material.

As a conclusion, the authors succeeded in writing a very interesting book from which the physicists and electronic engineers will learn how to make and how to use the surface wave filters. *L. Goraş*

RABINER L.R. GOLD B., *Theory and Application of Digital Signal Processing.* Prentice Hall, Inc. Englewood Cliffs, New Jersey, 1975.

Just ten years ago appeared a book written by two research workers of Lincoln Laboratory of Massachusetts Institute of Technology[1]). The book was devoted to open the way for a new discipline : the digital processing of signals. This pioneer work was stimulated, in turn, by two important facts : the large scale integration of solid state devices (the sixties) and the appearance of Cooley-Tuckey's algorithm (1965), concerning fast Fourier transform (FFT).

The present book (which keeps the name of one of the authors B. Gold) represents a reference book, with a large vision, devoted to improve both didactic and scientific experience, accumulated after the appearance of the first book.

As a general structure, the sequence of the chapters reflects a logical transition from abstract to concreet : from mathematics (theory of FFT and fast convolution as part of linear discrete-time systems) to physics (discussions on the pecularities of different filters in their impact with spectrum analysis) then to technology (hardware implementation of the integrated general purposes elements in the building of digital information processing devices as : filters, FFT computers and fast programmable microprocessors) and, finally, to the practical applications in the all-days life (speech study and radar improvements).

Strictly speaking of the form, we are concerned with a happy agreement of the physical reality with the involved mathematical reality. It is carefully emphasised the inherent limitations of this agreement and, by the way, let us remember a brilliant quandary of Einstein. Once. he said : „As long as mathematics is exactly, it doesn't describe reality and as long as it does, it isn't exactly". Naturally, we must understand here not the pessimistic but the gnoseological meaning of the idea.

Finally, from the content point of view, one can say we are concerned with an exhaustive up-to-date treatment of the problems by a critical and sincere weighing of all pros and cons. The encyclopedic content, the concision, the clarity and the excellent graphics make the book a good instrument in research work and a way of spreading the advanced ideas in the structures of high school teaching. *P. Pogângeanu*

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[1]) Gold D., Rader M. Ch., *Digital Processing of Signals* (with a subsidiary contribution of the authors : Th. G. Stokham Jr. — Ch. 7 and A.V. Oppenheim — Ch. 8) Mc Graw Hill Book Co., N.Y., 1969.

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— Designed for : measuring and controlling the level of filling materials (granular, powder-holding liquid) in reservoirs, bunkers, silos etc. between two pre-set limits.
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— Response time : less than a second.

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— Designed for : counting the heat quantity yielded by the thermal agent (overheated water) to the consumers.
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— Admissible eror : ± 1% for t = 20° ... 110°C.
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